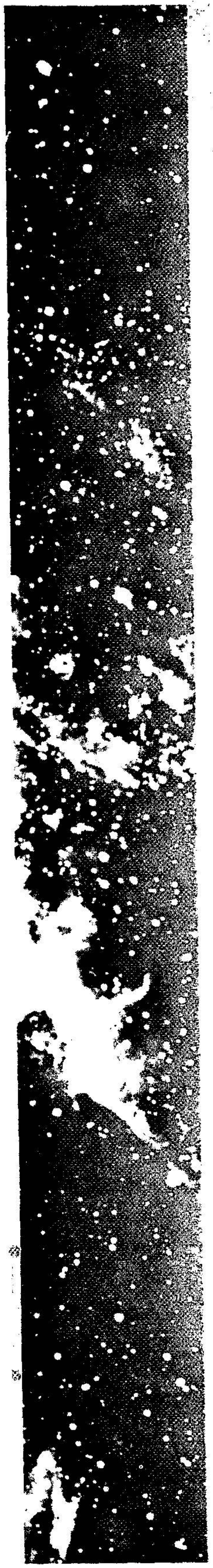
N.I.

CR-119803

20 FEBRUARY 1971

SPACE SYSTEMS



DEVELOPMENT OF CIRCUITRY FOR A MULTIKILOWATT TRANSMITTER FOR SPACE COMMUNICATIONS SATELLITES

Technical Report

Prepared For

GEORGE C. MARSHALL SPACE FLIGHT CENTER

NATIONAL AERONAUTICS AND SPACE ADMINSTRATION

HUNTSVILLE, ALABAMA

Under

CONTRACT NAS 8-24771

GENERAL ELECTRIC

20 February 1971

DEVELOPMENT OF CIRCUITRY FOR A MULTIKILOWATT TRANSMITTER FOR SPACE COMMUNICATIONS SATELLITES

TECHNICAL REPORT

Prepared for: GEORGE C. MARSHALL SPACE FLIGHT CENTER
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION
HUNTSVILLE, ALABAMA

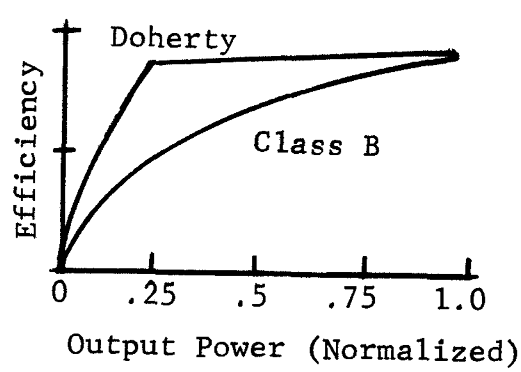
Contract No. NAS 8-24771

GENERAL ELECTRIC COMPANY
SPACE SYSTEMS ORGANIZATION
VALLEY FORGE SPACE CENTER
P.O. BOX 8555
PHILADELPHIA, PA. 19101

SYNOPSIS

This contract was directed at the development of a high power, high efficiency UHF transmitter which would serve as the basis for developing an optimum satellite transmitter for AM-TV broadcasting. The transmitter employs a highly unique UHF Doherty amplifier circuit adapted to 830 MHz (channel 73) where such a circuit had not been used previously. Special circuits and components were required for the grounded grid UHF amplifier configurations to provide the Doherty amplifier performance; included were radial-mode input and output cavities for each of the two tubes in the final amplifier, a special input power divider circuitry, and a dynamic biasing arrangement. Over the range tested with a stair-step TV modulated signal, the Doherty amplifier provided from 13% to 58% greater plate efficiency, depending on signal level, than an equivalent output Class B linear amplifier. For satellite operation where prime power is very costly, this improvement in efficiency is very advantageous.

The Doherty amplifier, originated in 1936, is ideal for a linear amplifier requirement. It utilizes two tubes, one to provide low power operation but which adds ever-increasing power as the signal magnitude approaches a maximum. The second tube reinforces the peak powers at the higher signal levels. The efficiency of the Doherty configuration compares with a Class B linear as follows:



Much of the signal power in an AM-TV picture signal is below the 50 percent output point, so the average TV efficiency improvement is seen to be quite substantial with the Doherty.

The Doherty was selected for development after evaluating a considerable number of high efficiency amplifier circuits and transmitting devices in an earlier contract. The UHF transmitter was expected to be more readily adaptable to near-future TV broadcast satellite systems than would microwave frequency transmitters which would require modifications of ground receiving sets.

The transmitter was designed specifically to have features which would be required in a space transmitter. These included monitor points and circuits to detect instantaneous faults and to remove them before they can cause tube damage, thermal control of tubes although water cooling replaced heat pipes in the tests, and a high efficiency planar triode based on an earlier research tube which provided superior UHF performance. The final Y2042 tubes were not available in the contract period and a lower-powered version, the Y1498, was used for developing the transmitter. The transmitter is designed for a 5 kW sync peak AM-TV output signal, which would be equivalent to an average power level of about 2 kW (CW).

Initial vacuum environment tests were performed on a section of coaxial line. A higher power level than available was necessary to produce a field of sufficient intensity to cause an actual electrical breakdown. Special sections of transmission line elements are available for testing when a higher power UHF source is available. The Doherty amplifier, itself, will be an ideal RF high-power source when the Y2042 tube becomes available.

The resulting Doherty amplifier is recommended for two general types of systems: (1) an AM-TV satellite requiring from a few hundred to several thousand watts of RF output in the UHF region and (2) a communications system requiring more than one channel in a common linear amplifier. The latter can profit from the developed transmitter because the triode amplifier, which is quite linear, reduces intermodulation distortion while the Doherty amplifier has a high average efficiency for all except very low overall signal levels.

TABLE OF CONTENTS

SECTION 3 (Continued)

ILLUSTRATIONS

TABLES

SECTION 1

BACKGROUND AND OBJECTIVES

1.1 PURPOSE OF THE MULTIKILOWATT TRANSMITTER PROGRAM

The ultimate objective of the overall Multikilowatt Transmitter Program is to place a high-power communication satellite transmitter in space in the early 1970's. The transmitter type to evolve will be coordinated with mission requirements established prior to the start of a prorotype development. In this contract, the mission is considered to be a direct or semi-direct TV broadcast function with conventional TV ground receivers, possibly using improved ground antennas but operating without receiver modification. The basic results of the study, however, will be applicable over a broad spectrum of missions.

1.2 PREVIOUS STUDIES

An earlier contract (References 1 and 2) examined key systems, subsystems, and components to identify optimum high-power transmitters for space satellite applications, and included investigations of specific critical problem areas. The study provided the parameters for high power space transmitter designs in the 50 watt to 20 kW range, and indicated techniques to implement thermal control measures and methods for preventing electrical breakdowns in the space environment. A recommendation leading to the present contract was to develop a high efficiency linear amplifier, specifically using the Doherty circuit for potential use in a UHF AM TV transmitter, and to experimentally evaluate the effects of the space environment on RF components that would be used in a multikilowatt transmitter in space communications and broadcast satellites.

1.3 OBJECTIVES OF THE PRESENT CONTRACT

The two major objectives of the present contract are to breadboard a 5 kW TV transmitter based on a high efficiency Doherty linear amplifier, and to verify

the ability of RF components to operate at high power in a vacuum environment without electrical breakdowns. These objectives involve breadboarding a linear driver and a 5 kW Doherty amplifier for a UHF AM-TV signal, an aural FM amplifier with a 500 Watt output, the necessary waveguide components to demonstrate operation of the transmitter, controlled carrier and energy storage circuits as required for AM-TV operation, and the protective circuits to prevent catastrophic failures should some electrical fault, temporary or permanent, occur.

The result of the contract is a breadboard transmitter that can be the basis for developing a space qualified transmitter for a UHF AM TV space broadcast or other communication mission. In addition, the critical interfaces, including the environment, thermal control, and dc power, are defined to permit the continuation of the program into a space qualifiable prototype transmitter.

SECTION 2

PROGRAM APPROACH

2.1 PROGRAM PLAN

An eight task program was formulated to accomplish the objectives set forth for the study of Circuitry Development for a Multikilowatt Transmitter for Space Communications Satellites. The tasks include the following, which are shown in the program plan block diagram of Figure 2-1:

TASK		TITLE
1		Transmitter System Design
2	a)	Driver (Visual Channel)
	b)	Doherty High Efficiency Amplifier
3		Aural Channel Amplifier
4		High Power RF Components; VSB Filter
5		Protective Circuitry
6		Controlled Carrier Circuit (Includes Energy Storage Filter)
7		RF Components Environmental Testing
8		Transmitter Tests

The purposes and objectives of each of the tasks are given below:

Task 1. Transmitter System Design

This task defines the approach to subsystem and component implementation. Trade-offs among the approaches to implementation of this particular transmitter were made with consideration for the basic circuit design, component availability and anticipated performance, and future applicability as spacecraft hardware. Items of particular concern were the transmission line, tube types, circuit characteristics, and initial mechanical design.

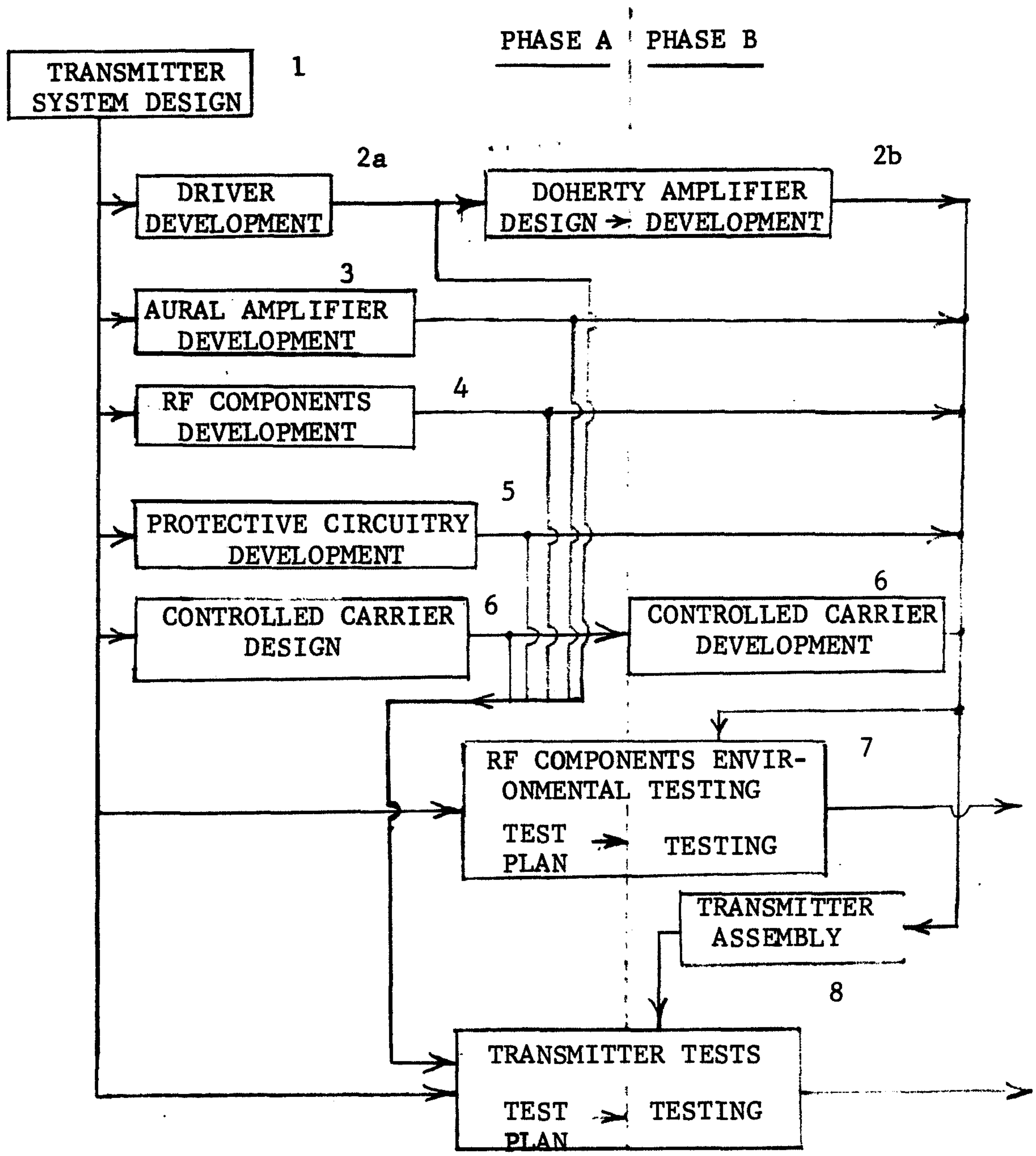


FIGURE 2-1. PROGRAM PLAN FOR MULTIKILOWATT TRANSMITTER CIRCUIT DEVELOPMENT

Task 2a and 2b. Visual Chain Amplifiers

This task involved the design, breadboard and component-test of the driver and the visual rf amplifier as required for implementation of the 5 kW (nominal sync peak level) TV transmitter. Design and development were to be accomplished for the high efficiency final visual rf amplifier (Doherty circuit) and the linear driver.

Task 3. Aural Chain Amplifiers

This task was to design, breadboard, and component-test the power amplifier for the 500 watt (nominal CW rating) aural FM channel for the 5 kW TV transmitter. The design of the final amplifier of this chain could serve as the model for portions of the visual chain Doherty amplifier.

Task 4. RF Components

The task included the design, breadboard, and component-test of rf components required for implementation of the transmitter breadboard; typical components are vestigial sideband filter, color subcarrier image filter, and power monitors.

Task 5. Monitoring and Protective Circuits

The elements required for transmitter protection and performance monitoring were to be included in transmitter breadboard. Typical circuit elements to be provided are crowbars, power monitors, and fault sensors.

Task 6. Controlled Carrier Design

A controlled carrier circuit suitable for inclusion in the breadboard TV transmitter was the objective of this task. This circuit permits the most efficient use of dc power by varying the average carrier level to adapt to the average modulation dc power requirements. An L-C energy storage filter was also to be included.

Task 7. High Power RF Component Environment Test

This task was to investigate experimentally the problem of electrical breakdowns in high power rf transmission line components in a space-simulated environment. A test plan was to be formulated to provide calibration data on multipactor and ion arc breakdown modes, and to show the ability of the transmitter's waveguide components to operate without failure. Test equipment was then to be assembled, and selected components tested at high power in a vacuum environment. Means for correcting breakdown situations would result as circumstances dictate.

Task 8. Transmitter Performance Tests

A test plan for testing the transmitter system comprised of the elements developed in the tasks above was to be formulated. The breadboard was to be assembled and complete system tests were to be performed in the later part of the contract, covering system performance and TV quality.

The outputs of the contract are:

1. Breadboard transmitter, including all components indicated as deliverable in the overall block diagram of Figure 2-2;
2. A final report which includes:
   - test data on the transmitter for a TV application
   - test data on the performance of high power rf components in a space simulated environment.

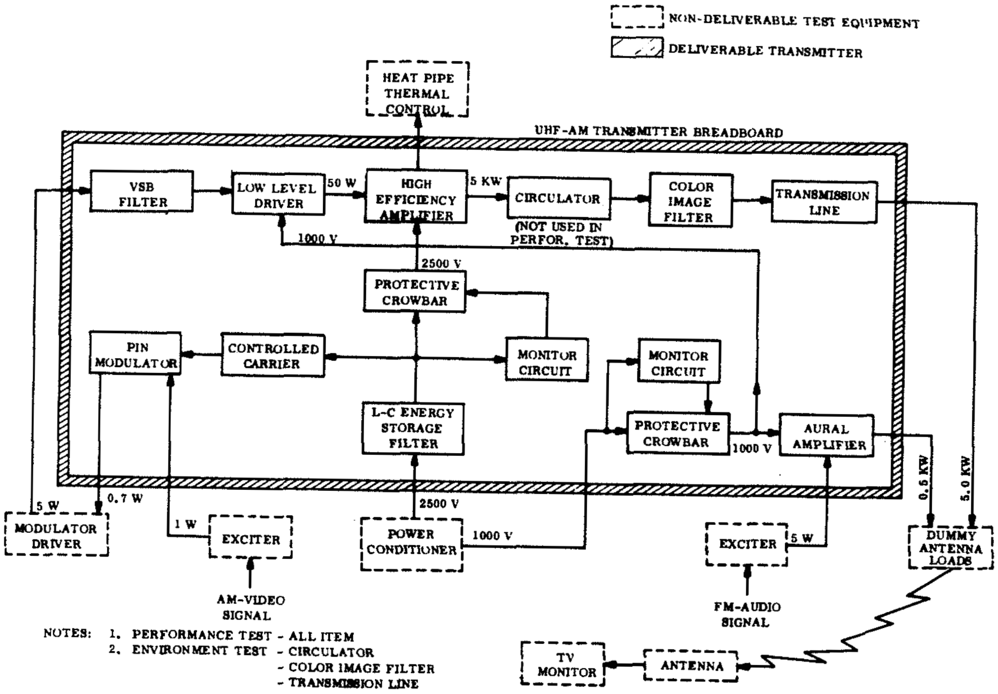


Figure 2-2. Block Diagram of Transmitter and Test Equipment

2.2 SCOPE OF EFFORT

The following provides a general indication of the scope for each of the tasks.

2.2.1 Transmitter System Design - Task 1

This task was the system definition phase of the program where the basic design of the transmitter is set, including the selection of major components. The reasons behind the selections were reviewed and consideration given to new developments which had taken place since the previous studies were completed. Then an approach to each required subsystem was defined in general terms, and specifications prepared for each.

The items relating to the various subsystems as set forth in the other tasks include

the following:

- Over-all detailed requirements, based on contract,
- RF Components to be used, including requirements for the vestigial sideband filter and color subcarrier image notch filter, and a comparison of waveguide and coaxial components,
- Tube Selection,
- Doherty circuitry as derived from the 30 MHz Simulator; also other amplifier circuits,
- General mechanical configuration of the transmitter and supporting subsystems, including thermal, power, and measuring equipment,
- Requirements for all protective circuitry, and an indication of circuitry suitable for the transmitter,
- Design requirements for a Controlled Carrier Circuit.

The results were used directly in defining the efforts in other tasks, and are included in the task discussions.

2.2.2 Visual Channel Amplifiers - Task 2

2.2.2-1 Visual Channel Drivers

This task was to design, fabricate, and test the Driver for the High Efficiency Doherty Power Amplifier and related associated circuitry. The driver amplifier is required to raise the power level of the external exciter (5 watts) to the nominal 100 watt level required for the Doherty amplifier input. The Driver must have essentially linear gain characteristics over the TV signal dynamic range, and adequate bandwidth to avoid excessive distortion of the television signal.

2.2.2-2 Doherty High Power Visual Channel Amplifier

This task was to design the final visual-channel amplifier based on the Doherty circuit. The design includes input and output cavity designs, dc circuit requirements, bias circuits, and interconnecting transmission lines at the input, output, and between the two stages. The high efficiency Doherty amplifier is required to provide a power output level of 5.0 kilowatts sync peak at the transmitter out-

put terminal. The amplifier must have essentially linear gain characteristics over the TV signal dynamic range and adequate bandwidth to avoid excessive distortion of the Television signal.

2.2.3 Aural Amplifiers - Task 3

This task was to design, fabricate, and test the aural channel output amplifier stage and associated circuitry. It was required to achieve a nominal 500 watt output level with a conventional FM aural signal, and be driven directly by a low power external exciter.

2.2.4 RF Components - Task 4

This task was to provide and test the specialized rf components required for proper functioning of the breadboard transmitter. The rf components interconnect the high-power amplifier outputs with the antenna and provide other functions such as vestigial sideband filtering and diplexing of the aural and visual signal outputs of the transmitter. Harmonic suppression was not considered necessary since harmonics would not affect fundamental frequency operation and they are normally low for a triode amplifier. Also, a harmonic filter would not add measurably to the technology since filter implementation techniques are well founded. The problem of potential harmonic filter breakdown is included in Task 7.

The specific waveguide components included in the system are a 3 dB hybrid, directional couplers for power monitoring, a color subcarrier image notch filter, waveguide-to-coax transitions to direct the rf power into dummy loads, and a low level vestigial sideband filter to be located at the input to the transmitter's driver stage.

2.2.5 Monitoring and Protective Circuitry - Task 5

This task was to design, fabricate, and test the necessary ancillary circuitry in the transmitter breadboard for protection of transmitter components from damage

under conditions of circuit malfunction or misadjustment, and for monitoring transmitter performance. The specific circuits include dc crowbars to cut off the power supplies if arcing or other dc breakdowns occur, and rf power measurements to turn the signal off if a high reflected power is detected indicating an rf fault. Control and logic circuitry are included.

2.2.6 Controlled Carrier Circuit Design - Task 6

This task involves the design of a "controlled carrier" modulator, or attenuator, for use with the high efficiency Doherty amplifier in the 5 kW AM-TV transmitter breadboard. This circuit permits the power supply and conditioner to be sized to the "average" transmitter power required for the TV signal, adjusting the carrier downward as the average modulation increases with a dark picture. The technique can reduce power supply and conditioner capacity requirements by as much as 40%, a highly significant saving in satellite and system weight and cost.

2.2.7 High Power RF Component Environment Tests - Task 7

A test plan was required to describe objectives, components, test techniques, and methods and basic limits to be utilized in testing rf components for high power multipacting and ionizing breakdown under high vacuum conditions as would be encountered in a space system. Emphasis was placed on multipactor breakdown, with an incidental consideration for avoidance of ionizing breakdown in the test system. A UHF rf source of 2.5 kW was to be used, employing the 500 W aural channel amplifier with a 20% duty cycle.

The items expected to be tested were:

Coaxial line - 3-1/8"

Waveguide - half-height WR975

Stepped coaxial line to identify multipacting conditions

Stepped waveguide section to identify multipacting conditions.

If circumstances permit, the following were also to be tested:

3-dB hybrid (sidewall) coupler

Color subcarrier image notch filter

2.2.8 Transmitter Tests - Task 8

This task was to establish the testing procedures for the Multikilowatt TV Transmitter, following EIA suggested methods whenever applicable, as included in EIA Standard RS-240.[3] Transmitter tests were then to demonstrate the ability of the multikilowatt transmitter to transmit high quality TV pictures with high efficiency. Simultaneous operation of aural and visual channels was required in the final procedures.

The tests were divided into three basic areas:

Tests of performance, including efficiency, power, and gains

Tests for TV quality factors, per EIA RS-240

Tests for TV performance with controlled carrier circuit included.

2.3 CONSTRAINTS

Constraints on the various tasks are determined from contract requirements, results of the System Design Study of Task 1, performance requirements such as EIA Standard RS-240, and available state-of-the-art components. These are summarized briefly in Appendix A, and pertinent factors are included in the appropriate technical descriptions of this report.

SECTION 3

RESULTS

This section summarizes the results of each of the tasks; detailed technical considerations and data are in Section 5.

3.1 TRANSMITTER SYSTEM DESIGN - TASK 1

3.1.1 Analysis of System Elements

The system design study considered the various elements of the transmitter system as they would influence overall performance. Included were components, circuits, and mechanical design; these were subsequent guidelines for other task developments. The overall task used the constraints of Appendix A to formulate the task direction; these constraints are based on various inputs from other programs as well as the EIA RS-240 TV Standard. Results of this task are distributed through the other tasks as appropriate.

3.1.2 The Transmitter System

A block diagram of the final transmitter is shown in Figure 3-1 while Figure 3-2 shows the unit (prior to adding the aural stage). The visual amplifier chain consists of the ML-8536 grounded-grid driver stage rated at 125 watts nominal sync peak output, and the Y1498 Doherty Amplifier (rated at 5 kw nominal sync peak output with future Y2042 tubes). The driver stage connects to the Doherty Amplifier through a power spliter/phase shift network which provides the appropriate drive signal levels to the two grounded-grid Doherty amplifier stages. WR975 half-height waveguide was selected as the transmission line; waveguide to coaxial transitions are included so that readily available dummy antenna loads can be used to absorb the amplifier RF outputs.

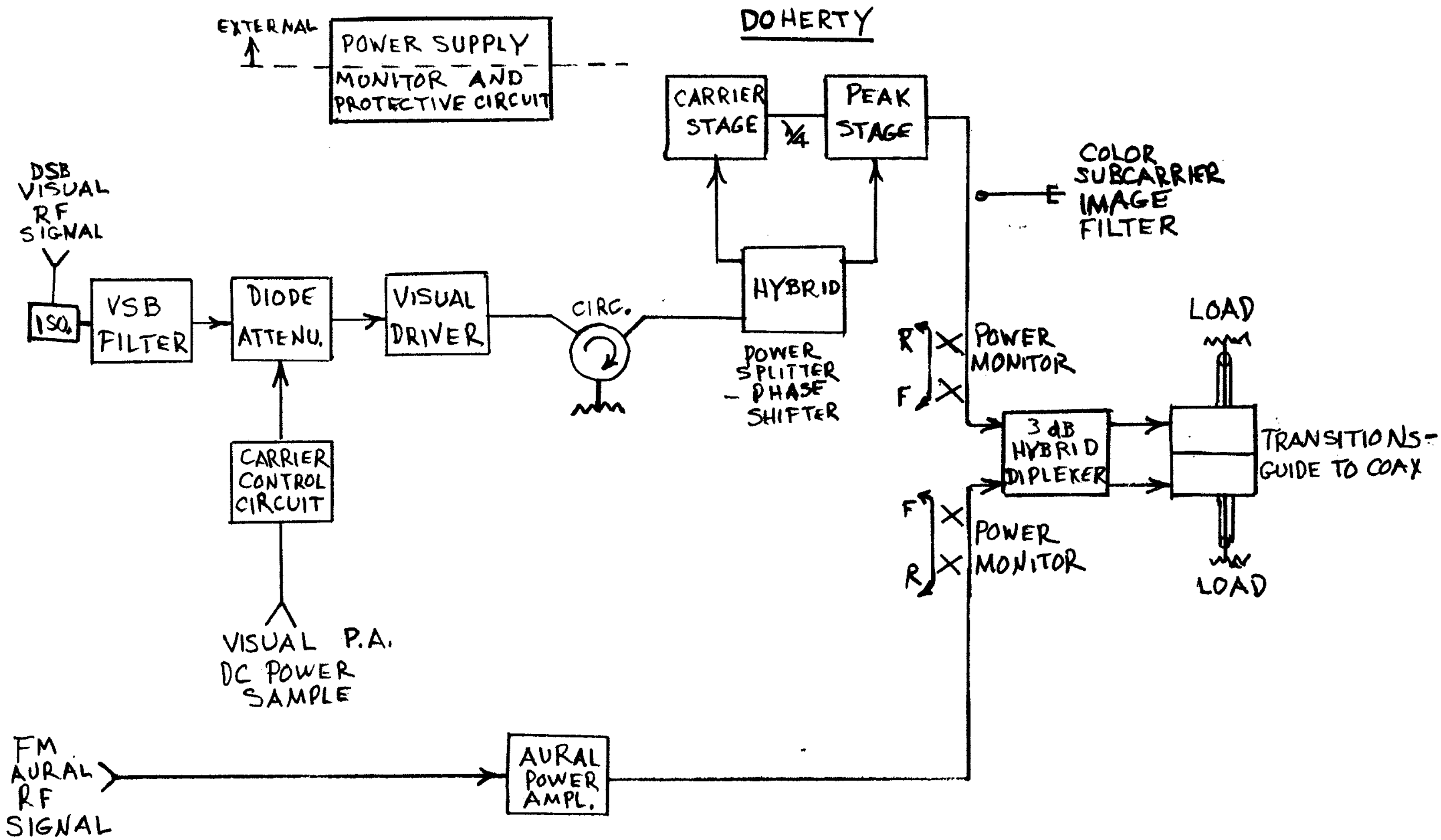


FIGURE 3-1. BLOCK DIAGRAM OF BREADBOARD TRANSMITTER

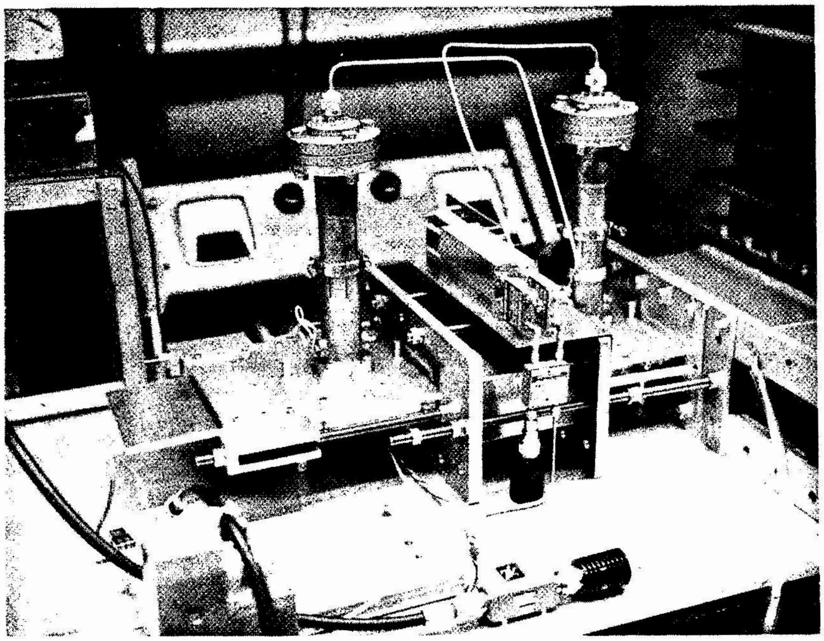


FIGURE 3-2. BREADBOARD TRANSMITTER

In the visual channel, low level VSB filtering at the input to the driver avoids the substantial inefficiency of an output filter, and weight and volume are greatly reduced. This approach necessitates a color subcarrier image filter at the output to suppress the intermodulation product at 3.58 MHz below the video carrier (from mixing of the video carrier and color subcarrier).

The aural stage in Figure 3-1 is connected to power monitors, and then combines with the visual rf signal through a 3-dB hybrid. The outputs to the two loads are equal power, identical signals for the breadboard test configuration of Figure 3-1. An additional 3-dB hybrid and suitable filters are used in an operational transmitter diplexer.[1,2]

Details on the transmitter assembly, testing, and performance are in Section 3.8. The amplifier demonstrated a definite efficiency improvement over a Class B linear stage, varying from 13% to 58% depending on signal level. Detailed experimental data is in Sections 3.2.2 and 3.8.

3.2 VISUAL CHANNEL AMPLIFIERS - TASK 2

3.2.1 Driver Stage - Task 2a

3.2.1-1 Design

The driver stage in the visual amplifier chain is a Class B linear amplifier, using a Machlett ML-8536 planar triode tube to obtain a nominal peak output of up to 125 watts to drive the high power Doherty amplifier. The driver is shown in Figure 3-3; a detailed diagram of the amplifier configuration is in Section 5.2.2. The grid is at dc ground as well as rf ground, thus assuring minimum rf feed-through and also a minimum multipactor likelihood in the output cavity. The input cathode circuit is resonated by a low impedance short-circuited transmission line of less than one quarter wavelength. The anode circuit is also a quarter-wave short-circuited coaxial

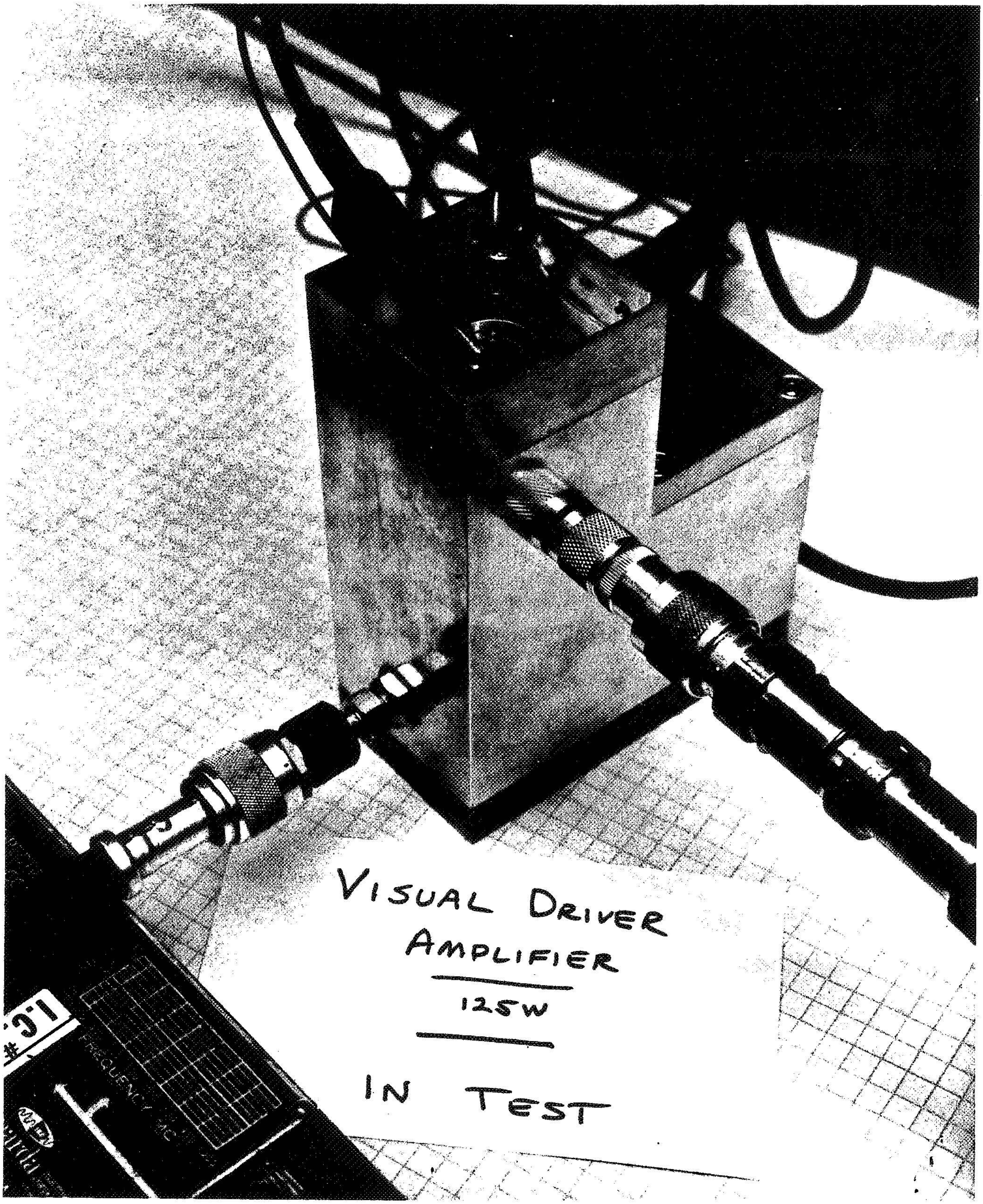


Figure 3-3. Visual Driver Amplifier

line; its impedance was chosen as a design compromise between a small diameter center conductor line to minimize the loaded Q and a large center conductor line diameter to minimize the temperature drop between the anode and the cavity surface. A second cavity is coupled to the first by an iris, thus forming a double-tuned circuit which will provide a wide bandwidth for minimum distortion. The tests in this report used a single cavity which was just adequate for test purposes but not preferred in a satellite system where it would not be accessable for adjustments.

3.2.1-2 Performance

The driver performance is as follows:

Output Power	117 watts max.
Drive Power	2.1 watts
Gain	18.1 dB
Efficiency	46%
Input VSWR	1.1
Load Impedance	4050 ohms
Plate Voltage	1200 volts
Grid Bias	-11.8 volts
Plate Current-Sync peak	227 ma
Grid Current-Sync peak	27 ma
Gain Suppression at Sync Peak	0.8 dB
Cavity with Tube:	
Loaded Q	69
3 dB bandwidth-single tuned	12 MHz

The cavities were cold tested. The resonant frequency of the cathode line was initially 875 MHz, which was decreased to 821 MHz by the addition of a tuning capacitor

between the case and cathode line near the cathode flange. The plate line resonated initially at 819 MHz, which was too close to the operating frequency for an added tuning adjustment to accommodate variations among tubes. The plate line impedance was raised slightly to reduce the resonant frequency, and a slug tuner was added to adjust the tuning precisely to the proper value. The above data was taken after tuning and loading was essentially at the optimum values.

The linearity capability of the driver can be seen in Figure 3-4, which shows a stair step signal following the sync peak signal (peak is negative in figure). Figure 3-5 supplements this by showing the actual differential gain measurement, indicating essentially no variation over the signal, and about 0.8 dB at sync peak realtive to the TV picture level signal.

3.2.2 Doherty Power Amplifier

3.2.2-1 Design

This task is to develop the Doherty high-efficiency visual-channel amplifier with a 5 kw sync peak output. Details on the principle of operation of the Doherty amplifier may be found in Reference 2 and in its references; the resulting amplifier circuit is in Figure 3-6, and the amplifier itself was shown in Figure 3-2. The circuit uses two Y1498 triode stages with the necessary intercoupling; the diagram shows the equivalent lumped circuit constants for the plate and cathode tuned circuits, although cavities, waveguides, and coaxial lines are used in the actual circuit.

This circuit introduces two principal design innovations into the basic Doherty amplifier. The first is the use of low impedance couplings, necessary at UHF, between the two plate cavities and also between the peak tube cavity and the antenna. Tests on a 30 MHz simulator[2] modeling the UHF cavity circuitry demonstrated the feasibility of low impedance couplings. The other major difference was the use of grounded grid circuitry which is the only feasible amplifier approach at UHF.

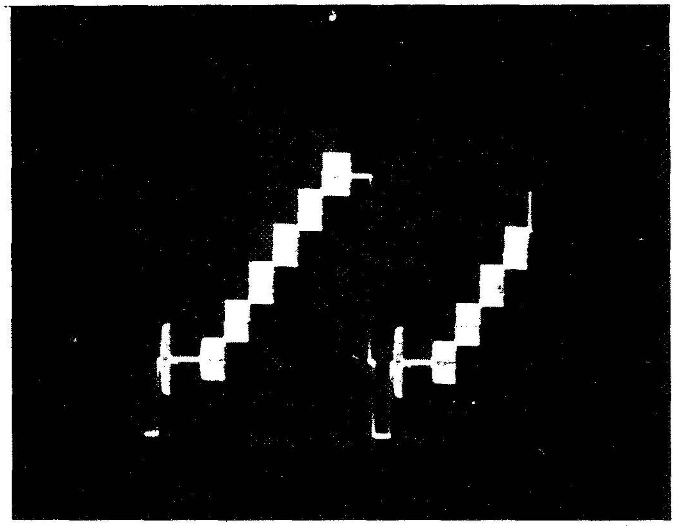


FIGURE 3-4. DRIVER LINEARITY TEST RESULT

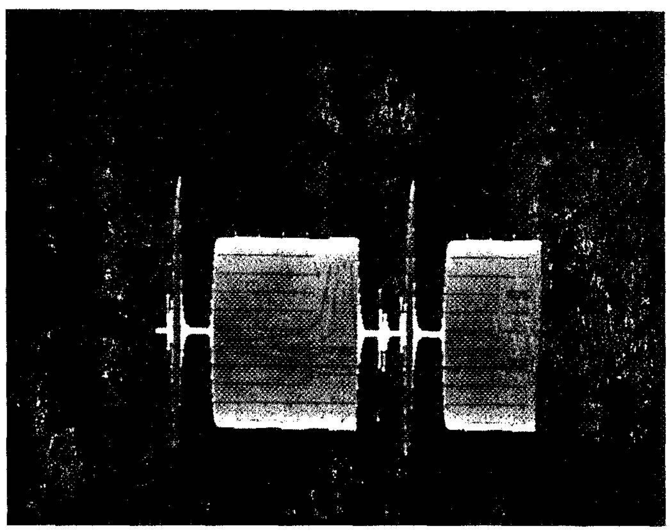


FIGURE 3-5. DRIVER DIFFERENTIAL GAIN MEASUREMENT

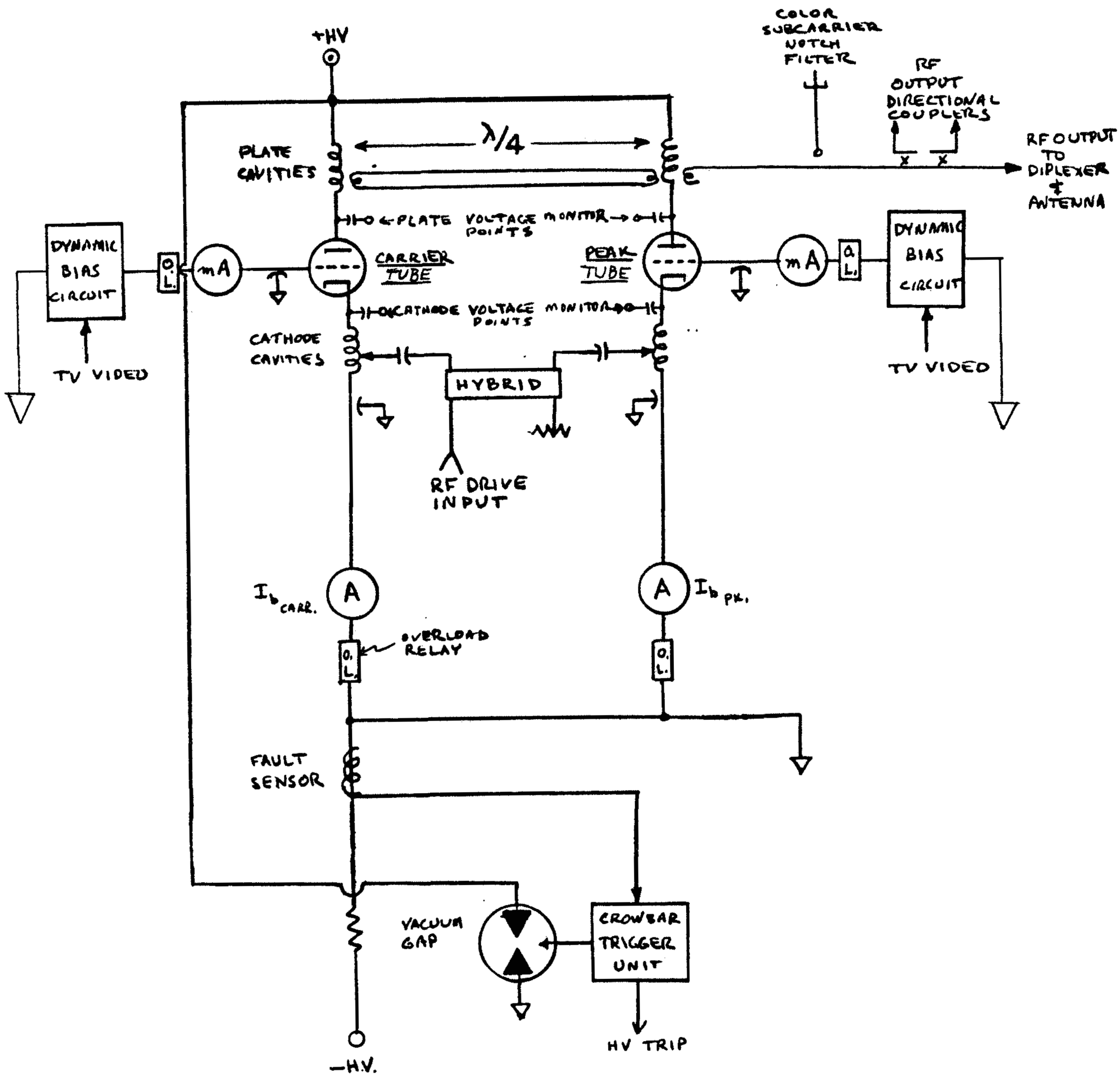


FIGURE 3-6. DOHERTY SCHEMATIC DIAGRAM

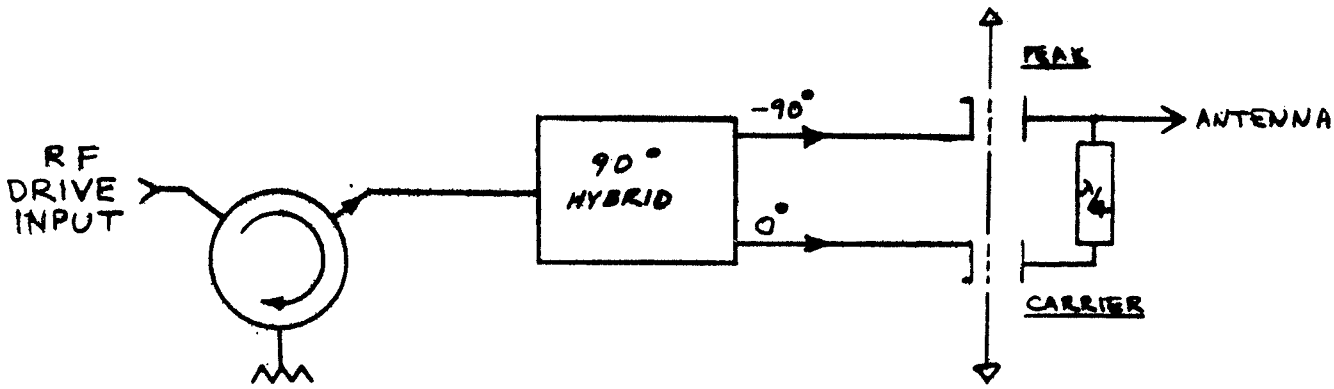
3.2.2-2 Assembly of Doherty Amplifier

The plate circuit cavities are rectangular to facilitate tuning and adjusting this somewhat complex amplifier. Output coupling is accomplished by means of an iris to a WR975 half height waveguide (discussed in Section 3.4.1), although some loop coupling was also added to the carrier stage because of iris geometry limitations. The stages were designed to use the high performance Y2042 triode, but only Y1498 lower power versions were available within the development time schedule. Quarter-wave coax input cavities (50 ohms) were used in both stages.

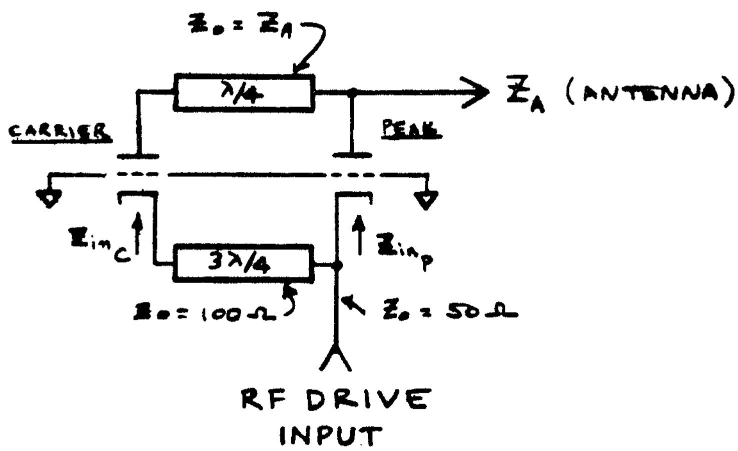
Tests on the 30 MHz simulator and analysis considerations indicated unique designs are required for the input power divider/phase shift circuit and for dynamic control of amplifier tube grid bias to suppress distortion levels and to achieve a high efficiency. The two approaches to input power division considered are indicated in Figure 3-7. The first approach uses a directional coupler type of power splitter which provides the required 90° phase shift between carrier tube and peak tube inputs. This was included in the transmitter circuit as developed. A second approach is to use a $3\lambda/4$ transmission line section between cathodes for impedance matching and phase shift control, with the input signal applied to the peak tube cathode. This approach could simplify the dynamic bias, as is described in Section 5.2.2-3.

The dynamic grid bias circuit varies the grid biases as a function of the instantaneous rf drive level. This is desirable in a Doherty amplifier for several reasons:

- the amount of drive on the carrier tube should cause saturation at carrier level (½ peak voltage)
- beyond this point, efficiency can be increased a little if the grid bias is changed to operate in the class C region
- as the drive is increased, the drive signal at each grid does not increase linearly due to changing input impedance as the triode grids begin to draw current
- by varying the bias accordingly, non-linearities can be reduced



HYBRID DIVIDER CIRCUIT



POWER SPLITTER CIRCUIT

Figure 3-7. Input Circuits for Doherty Amplifier

The dynamic bias circuit is shown in Figure 3-8, and the desirable variations of bias with signal level for the two stages are shown in Figure 3-9. The circuit is adaptable to both stages.

3.2.2-3 Doherty Test Results

Tests conducted with the Doherty Amplifier have demonstrated the substantial improvement which this circuit affords as a linear amplifier for AM television signals. Efficiencies obtained with the Doherty were as much as 1.58 times as great as that obtained with a class B amplifier operated under comparable conditions. The tests used a stairstep test signal similar to that shown previously in Figure 3-4. Details on the single-cavity Y1498 triode amplifier stage and the Doherty amplifier operation are presented separately in the following discussion.

Single Y1498 Cavity Amplifier

The two cavity amplifiers were the two stages used in the Doherty configuration. They were configured specifically for use with the Doherty Amplifier circuit to provide maximum flexibility in initial adjustment of the Doherty. Each uses the basic grounded-grid amplifier circuit of Figure 3-10 with a coaxial cathode (input) cavity and a radial line plate cavity which is iris coupled to the output waveguide line. A photograph of the amplifier is shown in Figure 3-11. Note the open end of the anode cavity which fits against the adjustable coupling iris and the adjustable side pieces which allow tuning of the cavity over a wide range to compensate for coupling iris reactance. An adjustable shorting plate (not shown) on the side opposite the iris is used for fine plate cavity tuning.

A series tuned coaxial circuit is designed to provide a resistive impedance at the input reference plane as the tube driving impedance varies with plate circuit loading. A high-isolation plate bypass capacitor was used to evaluate this type of

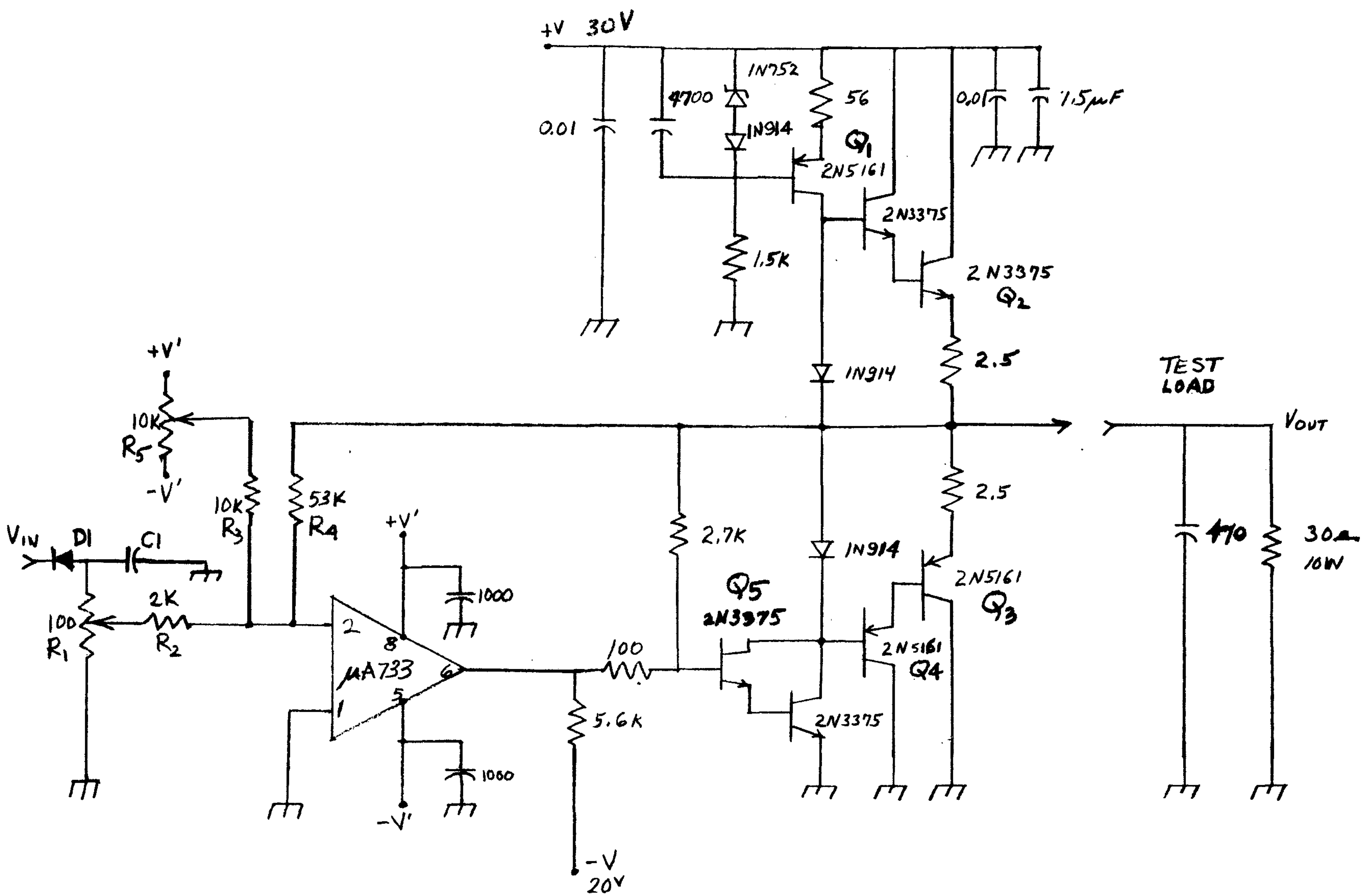


FIGURE 3-8. DYNAMIC BIAS SCHEMATIC FOR DOHERTY AMPLIFIER

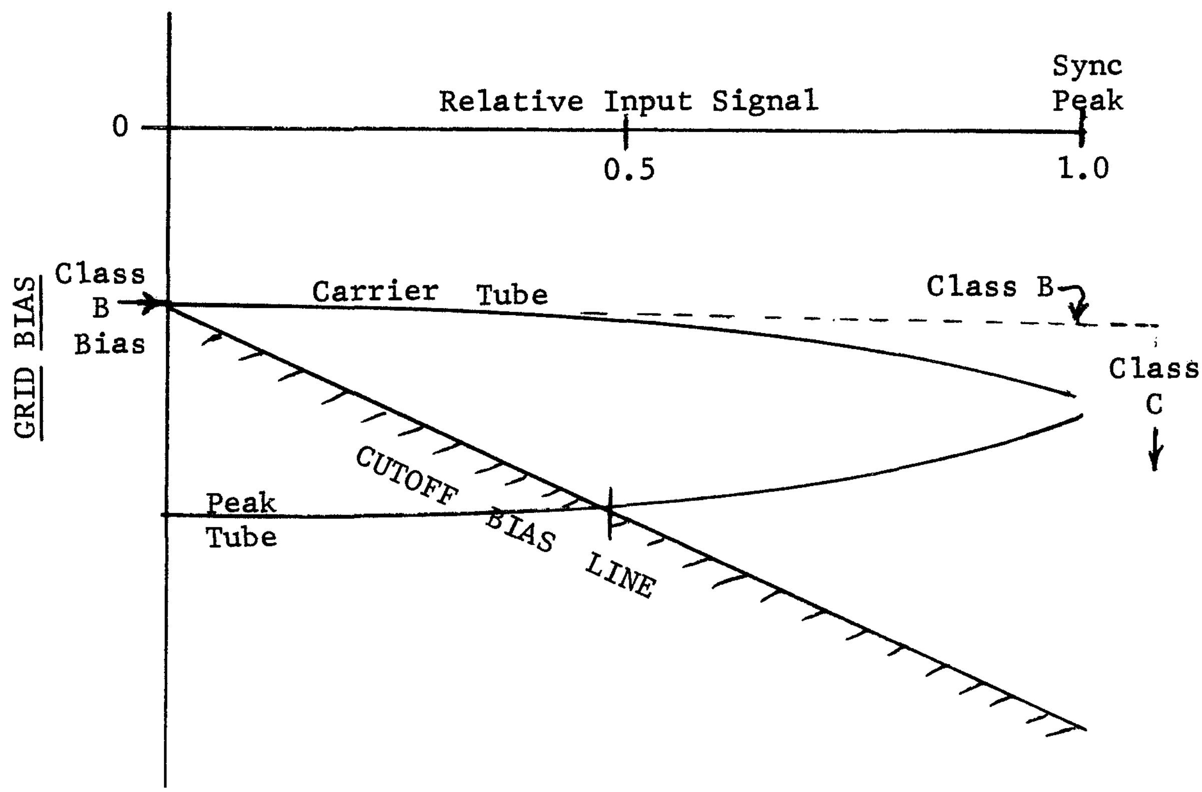


FIGURE 3-9

BIAS VARIATIONS ON DOHERTY AMPLIFIER TUBES

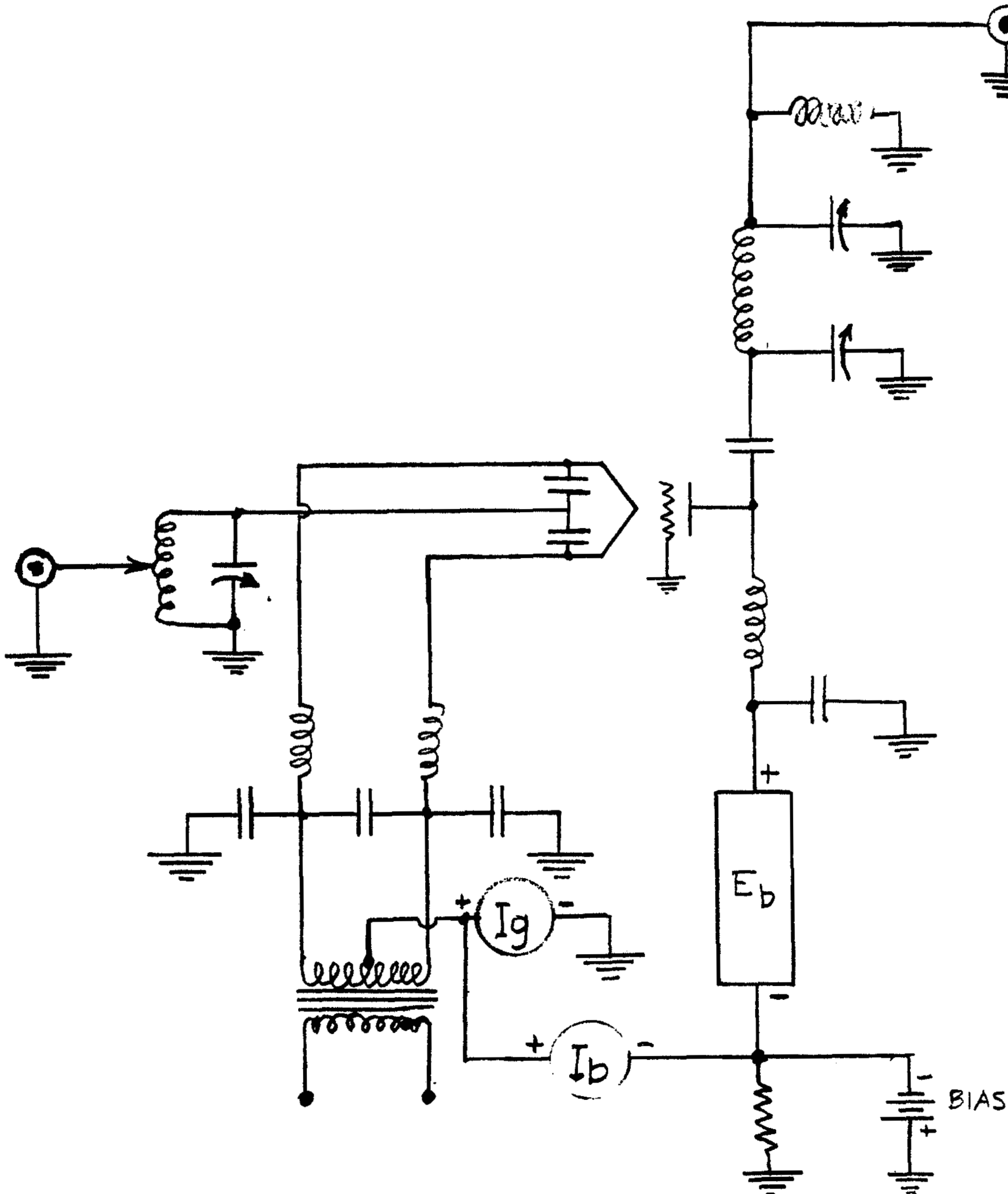


FIGURE 3-10. TYPICAL GROUNDED GRID AMPLIFIER CIRCUIT

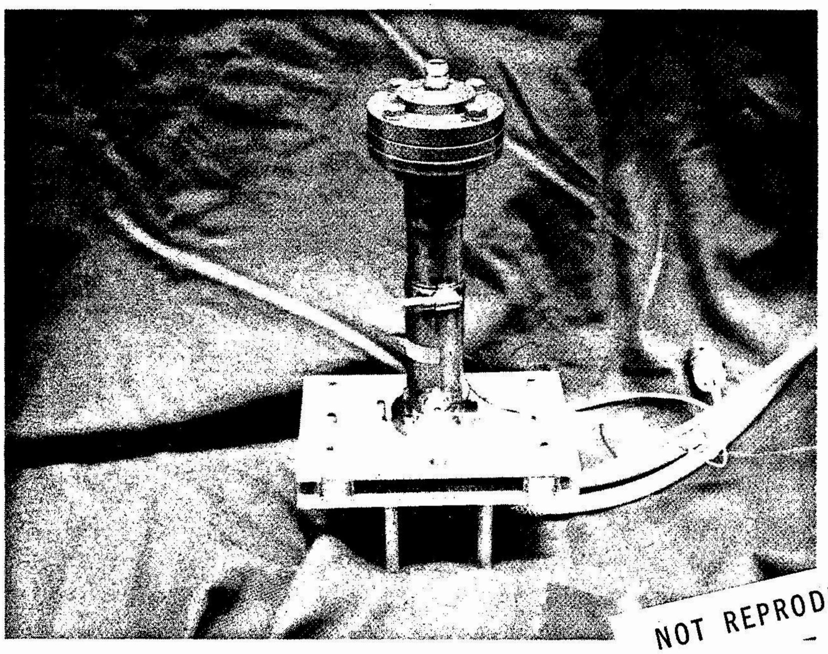


FIGURE 3-11. SINGLE STAGE OF THE 2-STAGE DOHERTY AMPLIFIER

component for subsequent use in dc grounded-plate, rf grounded-grid cavity designs where effective rf grounding of the grid and output-input decoupling are required.

Testing was conducted at moderate power levels while evaluating cavity rf characteristics to reduce the possibility of tube failure. This was desirable since probable grid overheating had been experienced in similar tubes used in the aural stage amplifier. Greatly improved thermal designs now being incorporated into the Y2042 tubes should allow full rated operation. A peripheral mode resonance which occurred at the intended operating frequency of 825.25 MHz made it necessary to test at a few MHz lower frequency and at a lower power to avoid excessive cavity losses and possible bypass capacitor failure. This is correctable in future amplifier designs by slight adjustments in capacitor dimensions.

The following test results represent normal rf loading for 2.5 kW sync peak operation at 2.5 kV plate voltage.

Test Results with #136 Y1498 Triode

Operating frequency	820 MHz
Plate Voltage	1000 V
Grid Bias	5.8 V
Plate Current	600 mA
Grid Current	240 mA
Output power (cw)	380 W
Gain	16.5 dB
Efficiency	63.3%
Bandwidth (3 dB)	9.0 MHz

Improvements in efficiency are expected in an operational amplifier due to decreased cavity losses which will result from the use of silver clad or plating of conducting surfaces rather than 6061-T Aluminum, elimination of a number of clamped and sliding

joints now incorporated for tuning purposes, and typical improvements in tube efficiency when operated at higher plate voltage (for the same plate load resistance).

Typical stairstep test results are shown in Figure 3-12. These waveforms are acceptable for television broadcasting with the normal gain-phase correction applied. (Note: The reduction of chroma level is due to filtering of the lower sideband and detecting with a double sideband demodulator.)

Complete Doherty Amplifier

The two Y1498 cavity amplifiers were incorporated into the Doherty circuit as illustrated in the photograph of Figure 3-13. Restrictions on power level and cavity loading were imposed due to the tube and plate-bypass peripheral mode resonance situations described above. However, the relative improvement in efficiency for two tubes combined in the Doherty circuit compared to a single tube operated under the same conditions as one tube of the Doherty amplifier is clearly evident. As an example:

	Class B Amplifier	Doherty Amplifier
Operating Frequency	820 MHz	820 MHz
Plate Voltage	1000 V	1000 V
Power Output (peak)	39 W	78 W
Efficiency (Stairstep signal)	14%	22%
Relative improvement using Stairstep signal		1.58

Television signal quality was unimpaired in the Doherty as indicated in Figure 3-14. The IRE filter was used on the TV waveform monitor so that the chroma signal is not present. Normal sync stretching and gain correction would be required as is customary in television broadcast practice.

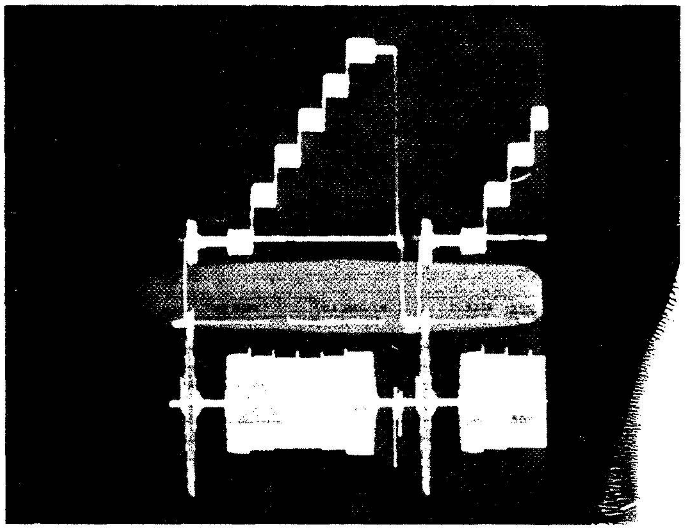


FIGURE 3-12. TV STAIRSTEP TEST OF SINGLE STAGE AMPLIFIER

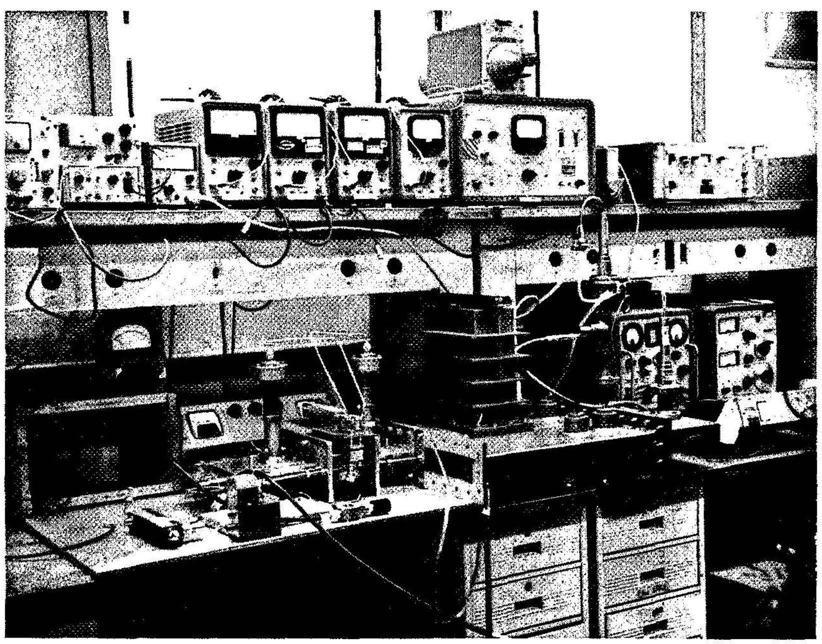


FIGURE 3-13. DOHERTY AMPLIFIER TEST INSTALLATION

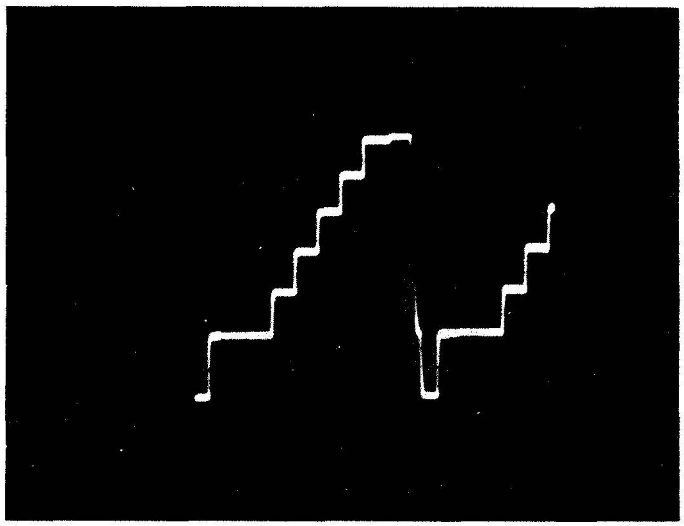


FIGURE 3-14. TV STAIRSTEP TEST OF DOHERTY AMPLIFIER

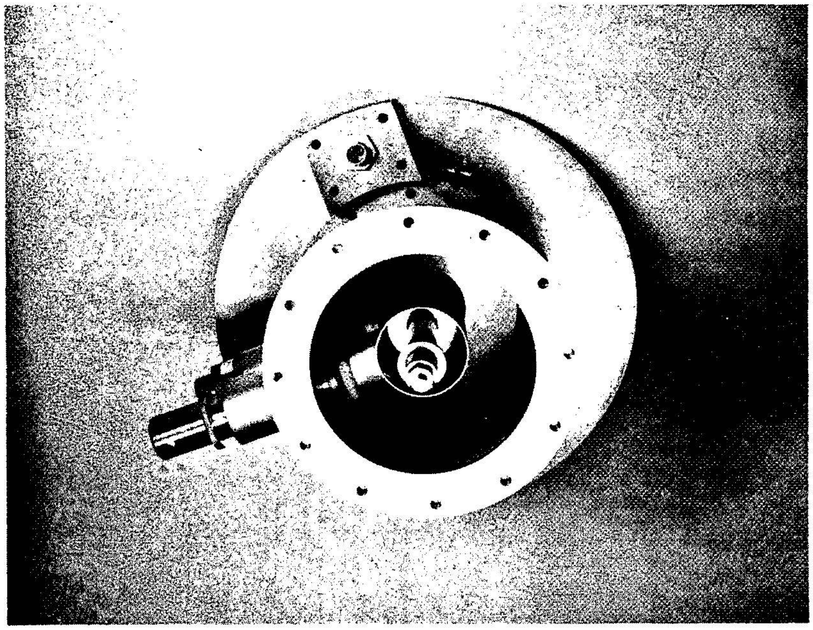
3.3 AURAL CHANNEL AMPLIFIER - TASK 3

3.3.1 Design

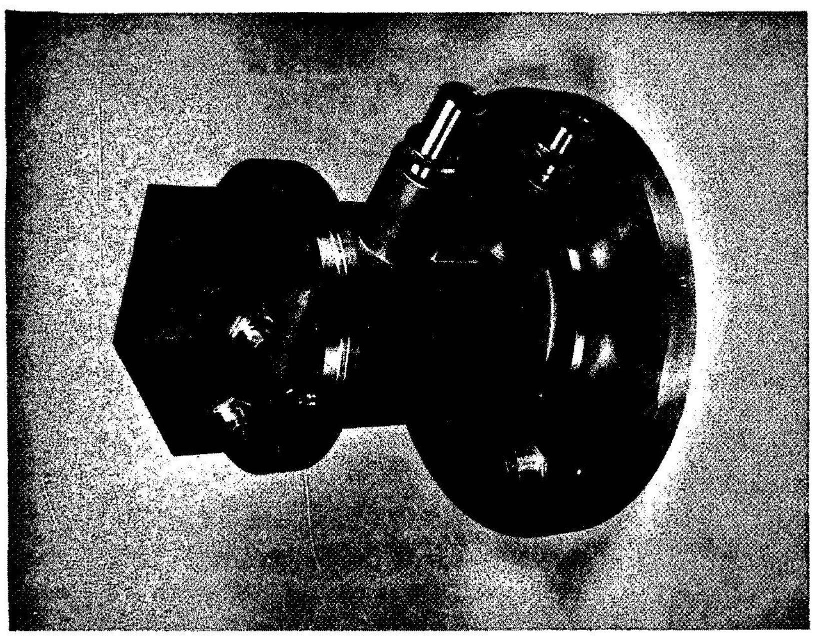
The transmitter requires a 500 watt FM aural channel amplifier to supplement the 5 kW AM visual channel amplifier chain in the transmitter. The amplifier uses a single Y1498 tube (adequate for this stage) in a Class C circuit. The amplifier has a three-quarter-wave coaxial input line cavity and a quarter-wave coaxial output cavity, using a dc grounded anode and rf grounded grid circuit. Figure 3-15 shows two views of the resulting amplifier; a detailed cross-section diagram of the amplifier is included in Section 5.3.

The simplest circuit applicable to planar triodes like the Y1498 tube at UHF is the rf grounded grid circuit employing coaxial resonators. The tube inter-electrode capacitances load the resonant transmission line sections, thus causing the lines to resonate at a frequency lower than that for which they are exactly a quarter wavelength (or $3\lambda/4$). To accommodate this effect, the cavity is designed with its physical length less than $\lambda/4$, just enough so the frequency is at the desired value. A $\lambda/4$ radial cavity has been selected for use between the plate and the grid, and a $3\lambda/4$ coaxial cavity between the grid and the cathode. The bandwidth reduction effect of the $3\lambda/4$ cavity (Sect. 5.2.2-2B) is not critical in a low Q input circuit.

Necessary dc operating voltages require that sufficient insulation be used, and the insulation should be compatible with the rf and dc circuits. The large value of dc blocking capacitance between grid and cathode and between cathode and anode can best be implemented with a folded cathode line configuration as shown schematically in Figure 3-16, or in detail in the amplifier cross-section in Figure 5-7 of Section 5.3. The folding approach permits the cathode/anode bypass to be located physically at the foldpoint of the folded line. The grid bypass is also conveniently located at the grid flange.



Input Cavity



Assembled

Figure 3-15. Aural Channel Amplifier

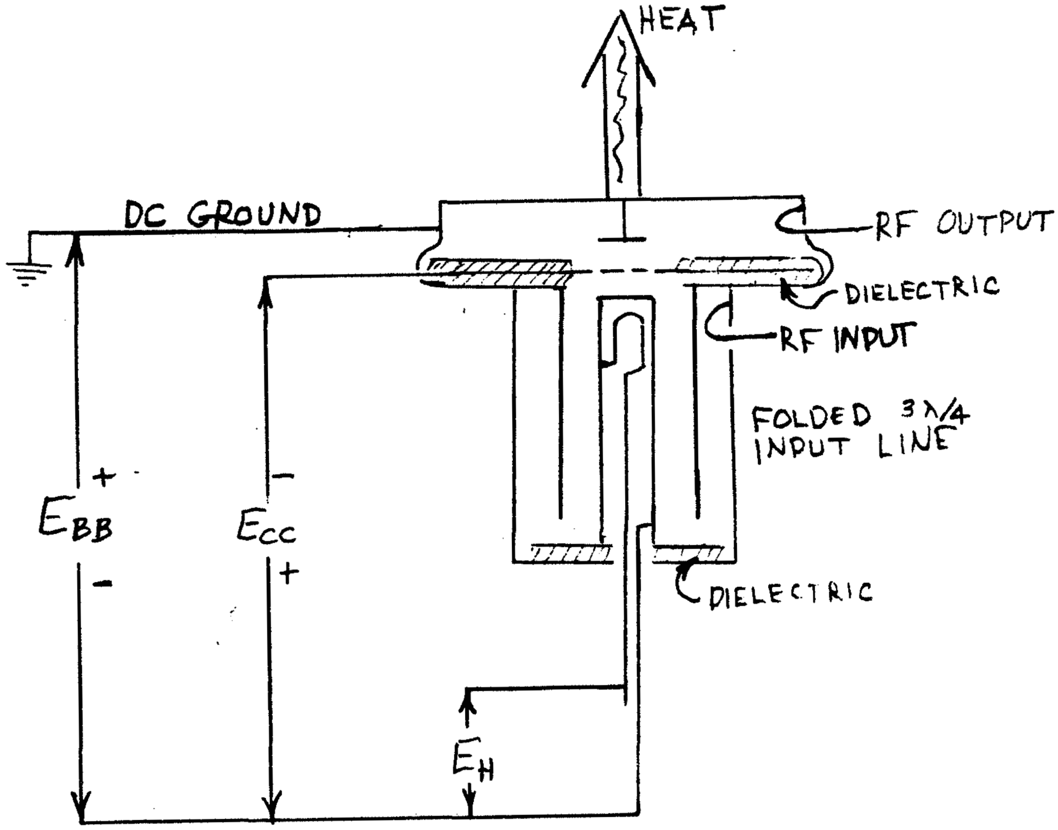


FIGURE 3-16. SCHEMATIC OF FOLDED CAVITY AURAL AMPLIFIER

3.3.2 Assembly

Certain features were included in designing the amplifier to meet specifications.

Some of these are:

1. The tube anode is dc grounded and thermally isolated from the cavity structure, permitting the cavity to operate at a lower temperature than the anode.

2. Some flexure is permitted between the tube and the two cavities so as to relieve stresses that might arise due to separate mountings for the cavity and the heat transport system attached to the anode.

3. The upper surface of the anode cavity provides a convenient mounting surface and heat transfer interface.

4. A copper flange soldered directly to the tube's grid contact surface provides a larger surface for grid heat flow to the cavity, which then conducts it to a final sink. The bypass capacitor insulation, through which the heat must flow, will also have a lower temperature differential if it has a large surface.

5. No dangerous voltages are accessible when the cavity is assembled.

6. The overall structure is compact and rigid.

3.3.3 Testing

The usual problems were encountered in the circuit, and one unusual one emerged. Both cavities had to be trimmed to meet the frequency requirements. Following this, however, an unusual amount of feed-through from the input cavity to the output cavity was observed. Since the tube was cold, the cause of the feed-through was difficult to identify. After extensive testing, the grid bypass capacitor was found to act as a $\lambda/4$ line, the rf passing through the insulator and around the grid flange (see Figure 3-16) to the anode cavity. This was corrected, after which tests were performed.

Initial tests on the amplifier used a coupling loop to extract power from the anode cavity. For the final configuration, an iris will be cut in the outer wall for coupling the anode coupling directly to the output waveguide. An aural channel amplifier as designed and fabricated was operated with the following characteristics:

Operating Frequency	829.75 MHz
Bandwidth - 3 dB	26.5 MHz

Plate Voltage	1550 volts
Operating Grid Bias	-15.5 volts
Plate current	0.76 amps
Grid current at full rf output	0.1 amp
Drive Power	40 watts nominal
RF power output	520 watts nominal
Plate efficiency	44%

Tests on a cold anode to grid cavity resonator were conducted to determine dissipative loading of the tube on the tank circuit. The Q of the test cavity was measured, and it was found to increase by factors of 2.7 to 5.3 (Q_u value as high as 900) after polishing and plating of tube electrodes and cavity parts.

3.4 RF COMPONENTS - TASK 4

Several RF components are required in a television transmitter to inter-connect the high power amplifier outputs with the antenna and to provide other functions such as filtering, power monitoring, diplexing of the aural and visual signal outputs, and harmonic suppression. Harmonic suppression is not considered necessary for this non-radiating breadboard system since harmonics will not appreciably affect fundamental operation. Possible electrical breakdown in harmonic filters is evaluated in Section 5.7.

3.4.1 Transmission Line

Results of previous studies[2] indicate that 3-1/8 inch 50-ohm coaxial line and half-height WR975 waveguide are reasonable choices for UHF space applications, with waveguide being the preferred approach from both the thermal and rf breakdown considerations. Very low impedance ridged-waveguide and coaxial lines offer promise of freedom from multipactor effects, but impedance matching requirements and higher losses

are penalties for their use. Half-height WR975 was thus selected as optimum. No unusual weight reduction measures were used since they would increase the program cost and stretch the schedule without influencing electrical performance. Low power and very short lines may be coaxial while the VSB filter can use stripline satisfactorily.

3.4.2 High Power RF Components

The components which were designed, fabricated, and tested in this task are:

- Color Subcarrier Image Rejection Filter
- Incident and Reflected Wave directional couplers (power monitoring)
- 3 dB Hybrid (borrowed - not a contract component)
- Dual Waveguide-to-Coaxial Transitions

The RF components for the breadboard transmitter are shown in a sketch in Figure 3-17. The requirements for these components are included in Appendix A; the equipment designed and fabricated to meet these requirements is discussed here. Photographs of the rf assembly are shown in Figure 3-18.

3.4.2-1 Color Notch Filter

The color subcarrier image notch filter used for the breadboard transmitter is a top-wall coupled single cavity type which conforms to current TV design practices and was found to be adequate(2). This filter is a high-Q type tuned to 821.67 MHz; a calculation has indicated that the Q should be of the order of 18,000, which is attainable with a WR975 dimensioned cavity. The calculated loss at the resonant frequency is then of the order of 0.1 dB, and rejection of the unwanted signal is 20 dB. The device is shown attached to the waveguide assembly in Figure 3-18; it was left unfolded for the breadboard design. Tests made on the device indicated it operates within the loss and rejection specifications; loss is less than 0.1 dB, rejection more than 20 dB, and VSWR of 1.23. The latter is sufficiently high to

ITEM	DESCRIPTION*
1.	Color image notch filter/ directional coupler assembly
2.	Hybrid, 3 dB sidewall coupler
3.	Adapter, dual, waveguide to 1-5/8" EIA coaxial line, 50-ohm

*Note-All waveguide connections are dual flange, 1/2-height WR975 waveguide.

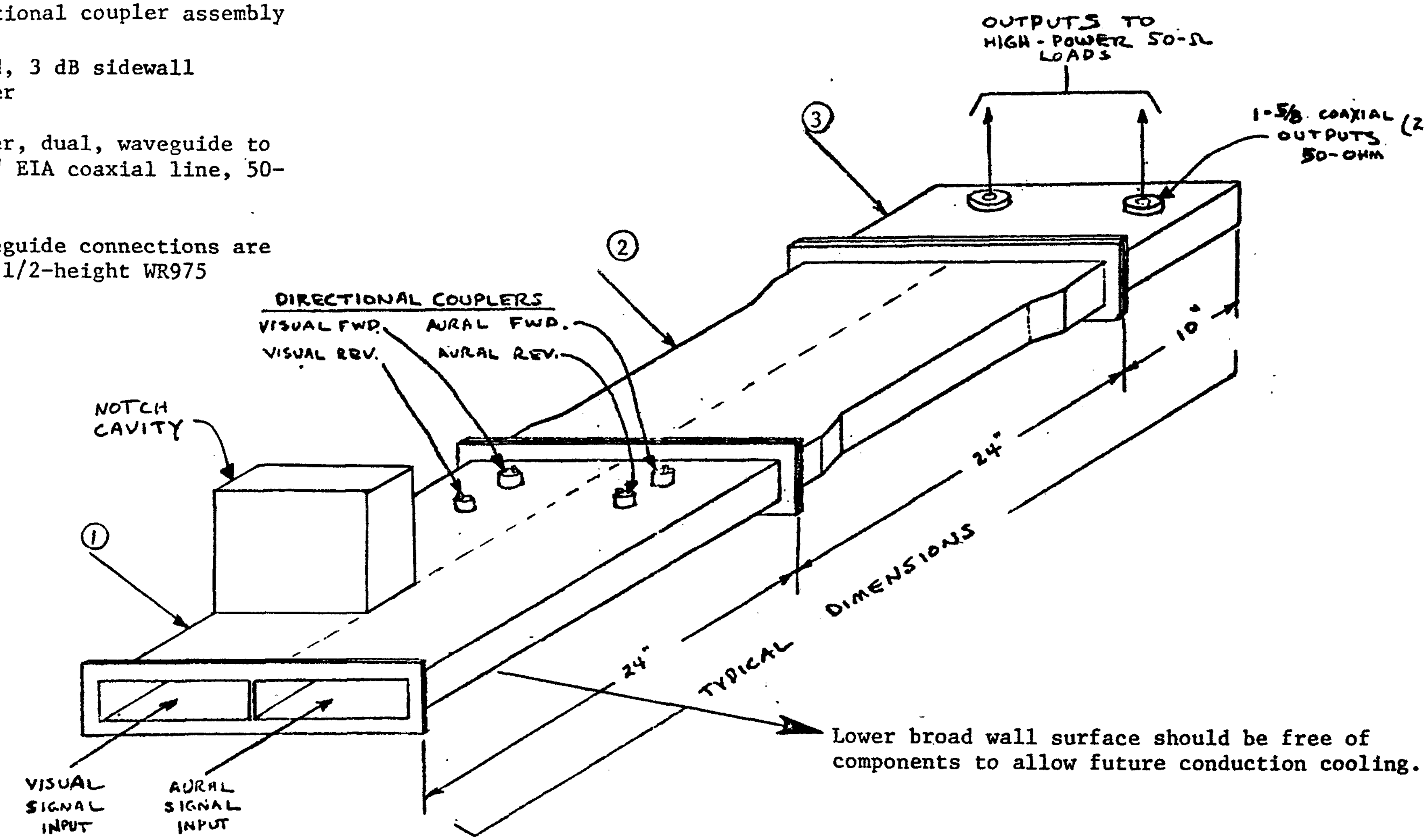


Figure 3-17. Waveguide Assembly for TV Transmitter Breadboard Setup

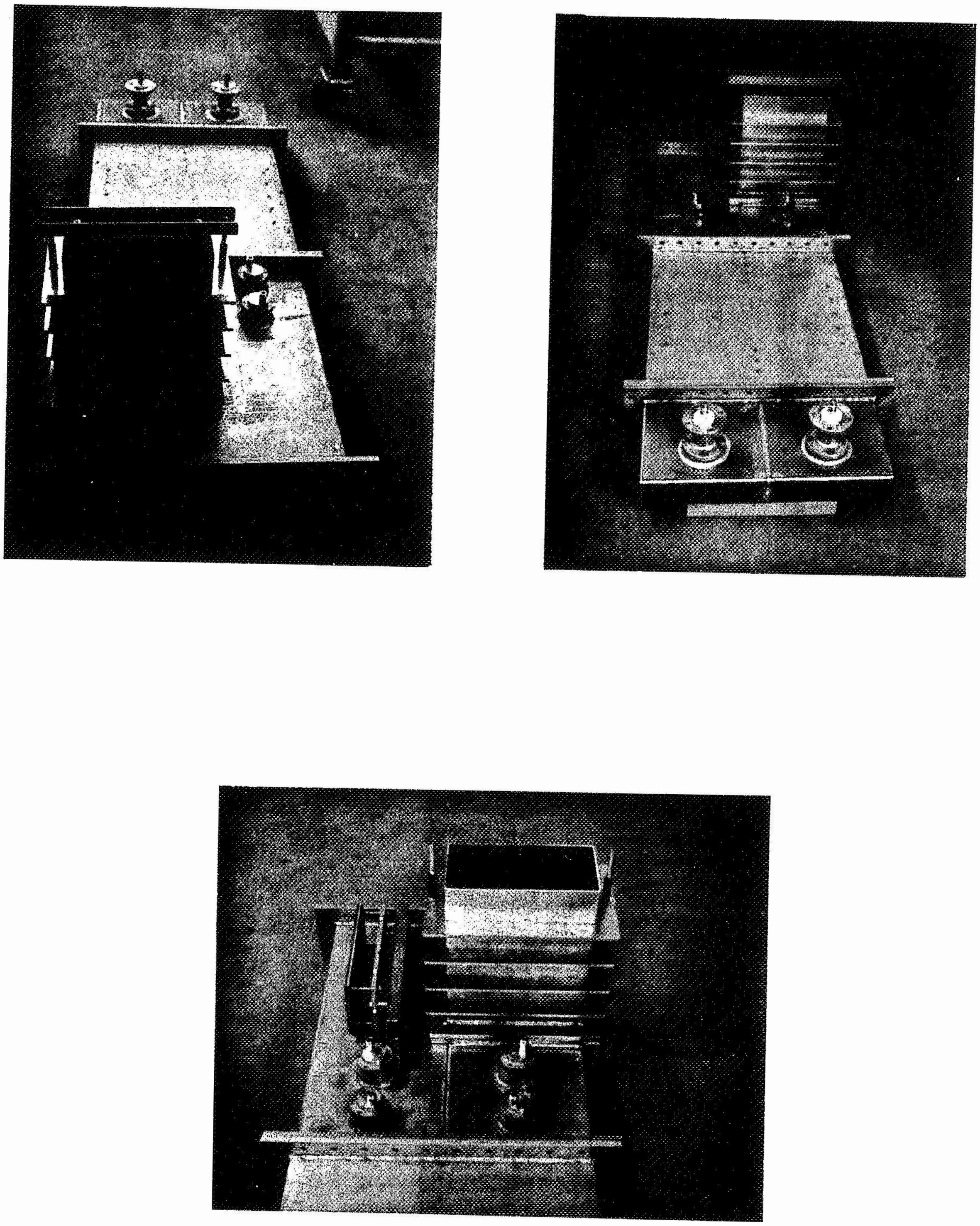


Figure 3-18. Fabricated Waveguide Assembly

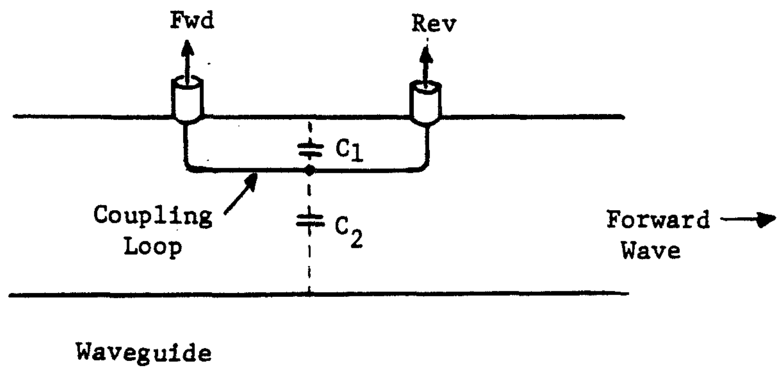
require additional tuning in the assembly. These data include the directional couplers discussed in the next section since the filter and directional couplers were assembled as a unit.

3.4.2-2 Directional Couplers

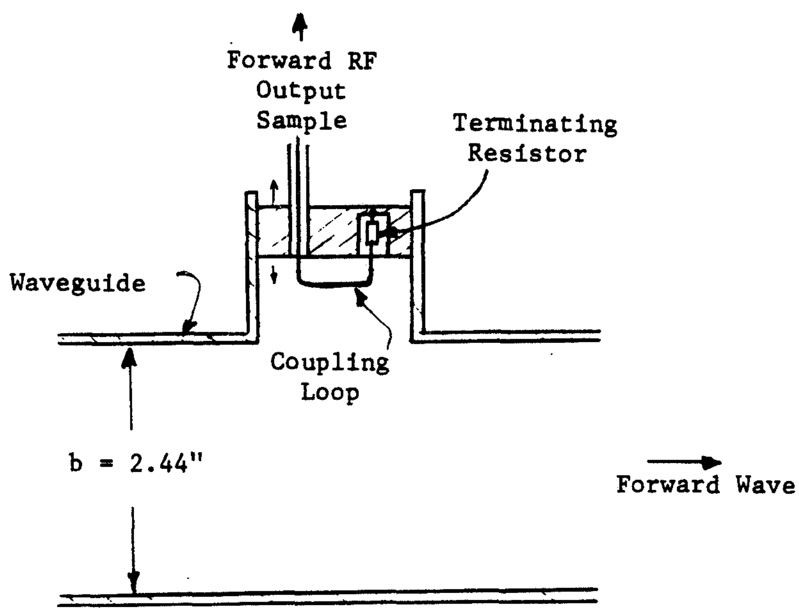
The reflectometer type directional couplers have adequate performance in the UHF band, and were chosen on the basis of compactness. The sensitivity of this type increases by 6 dB per octave, but the narrow band of interest for this breadboard makes this variation negligible. An illustration of the basic form of the coupler is shown in Figure 3-19. A small loop is introduced into the waveguide, which couples to both the magnetic and electric fields by virtue of its orientation, and is terminated by a coaxial output line of Z_o impedance. Two couplers are used in each channel, one each to monitor forward and reverse powers. These couplers are detailed in Section 5.4.1-2; measurements on the final couplers indicated -50 dB forward coupling, -40 dB reverse coupling, and 30 dB isolation, all of which are sufficient for this program. Insertion loss is considered negligible.

3.4.2-3 3-dB Hybrid

This component is included in the test because of its availability as a loan item from another program. The tests could be performed with the visual and aural output signals entering separate absorptive loads, but the 3 dB hybrid combiner will provice an indication of any load interaction effects which might occur with a diplexer in the system. Each of the two output waveguides thus will have half of the total aural and half of the total visual signal powers. The hybrid is included in the photographs of Figure 3-18. Its loss is negligible ($<$ 0.1 dB), coupling is 3.0 dB, isolation is 33 dB, and VSWR is 1.03 at band center.



(a) Basic Reflectometer Type Directional Coupler



(b) Arrangement of Couplers Used in the Breadboard Transmitter

FIGURE 3-19. DIRECTIONAL COUPLERS

3.4.2-4 Waveguide-to-Coax Transitions

Coaxial loads with 1-5/8" connectors where available for use in the tests, so two transitions interconnecting the waveguides and the 1-5/8" coax lines to the loads are included. The loss in a transition was measured to be less than 0.1 dB; each of these transitions can easily handle half the 5.5 kW peak power (aural plus visual) in each of the output waveguides of the RF system. The transitions are included in the waveguide assembly in Figure 3-18. Their VSWR was measured to be 1.03.

3.4.2-5 Assembly

With all components assembled, measurements indicated VSWR's, after additional tuning, of 1.10 in both the visual and aural channels; insertion loss is estimated to be about 0.1 dB. These data indicate a well performing output waveguide system which is entirely adequate for the Doherty transmitter.

3.4.3 Vestigial Sideband Filter

The response requirements of a vestigial sideband filter for television transmission, are shown in Figure 3-20; they include a 20 dB skirt drop-off in a 0.5 MHz interval. This skirt requirement can be met with a filter design using the phase sensitive properties of a 3 dB quadrature hybrid as shown in the block diagram of Figure 3-21.

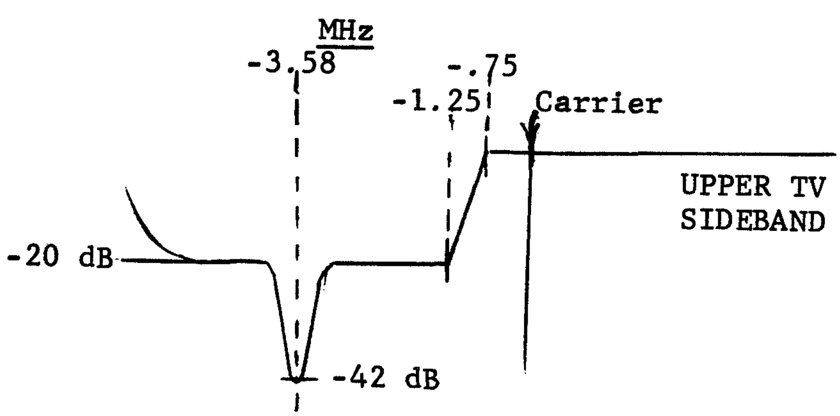


FIGURE 3-20. VESTIGIAL SIDEBAND FILTER RESPONSE SPECIFICATION

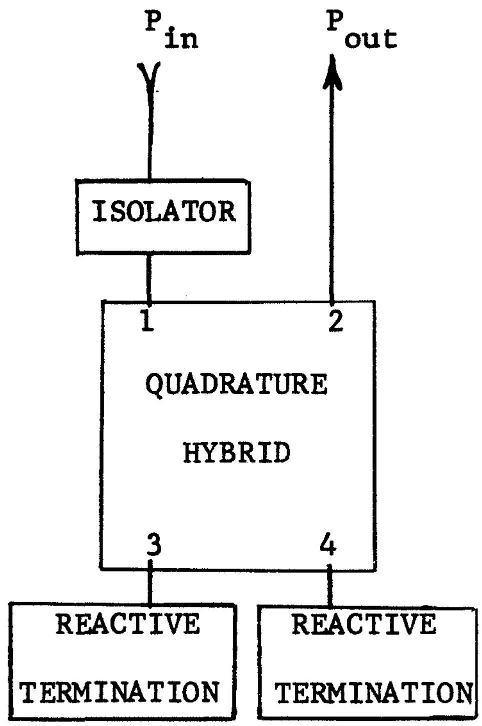


FIGURE 3-21. BLOCK DIAGRAM OF A VESTIGIAL SIDEBAND FILTER

The quadrature hybrid is sensitive to the relative phase and magnitude of the termination at each of its 3 dB ports. If the signals reflected from both terminations are in-phase, the hybrid will pass all of the incident energy. If the two reflected signals are 180° out of phase, then all the energy is reflected back to the source and there is no output. The vestigial sideband filter makes use of this characteristic to yield a bandstop filter response with very steep skirts. Since the phase of a resonator changes much more rapidly than its amplitude, a much steeper skirt is obtained from the phase type filter than can be obtained using conventional techniques.

The filter configuration fabricated for this program used one double-tuned and one single-tuned cavity as terminations on the hybrid; the filter block diagram is shown in Figure 3-21. Figure 3-22 illustrates the computed filter response for the selected configuration and includes the ideal response curve for comparison. The filter, which must be capable of dissipating up to 5 watts (i.e., half the maximum input power to the driver) is fabricated of stripline; Figure 3-23a shows the conductors before final assembly and Figure 3-23b displays the completed final assembly, the dimensions of which are 6" x 10" x 3/4". The ground planes are made of 1/8" aluminum plate while the intervening space is filled with four 1/8" polystyrene plates; the stripline structure employs one mil brass shim stock centered in the sandwich. In the space version, the ploystyrene will be replaced with PPO material.

Initial tests on the filter have shown the rejection band to be broader and the slope of the skirts to be less than predicted. Measurements were made on the resonator Q's and they were about 1000 as compared to a desired value 3 or 4 times this. The low Q's indicated the necessity for using silver plated stripline, and probably larger conductors, to obtain less loss. In addition, the use of double-tuned circuits for both terminations of Figure 3-21 should provide a steeper skirt drop-off.

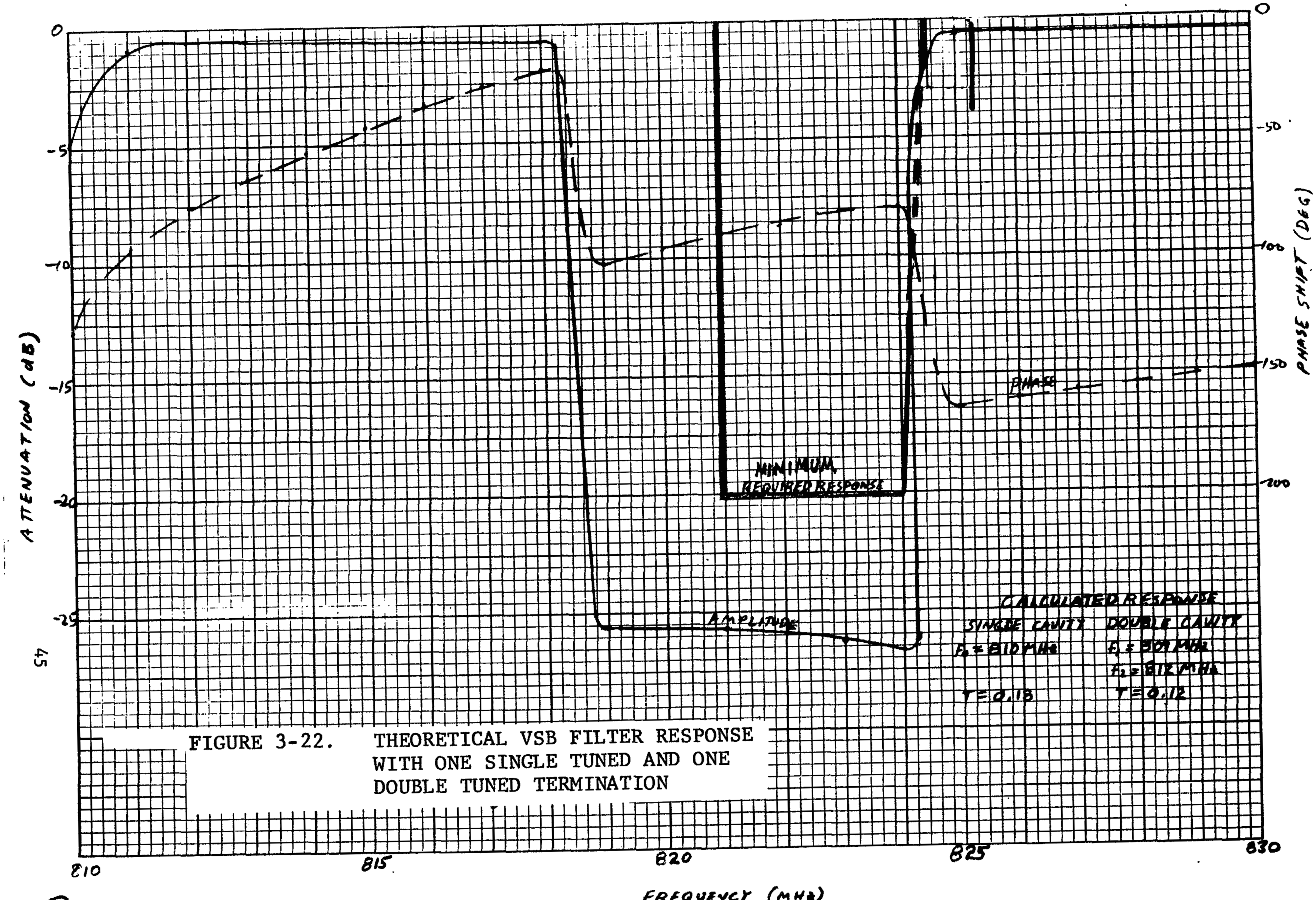
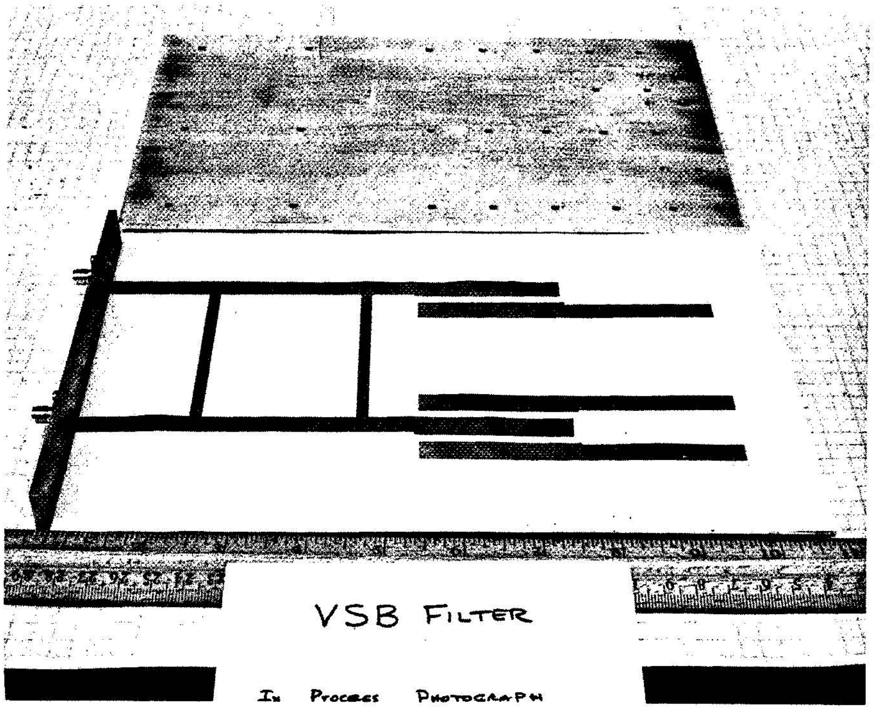
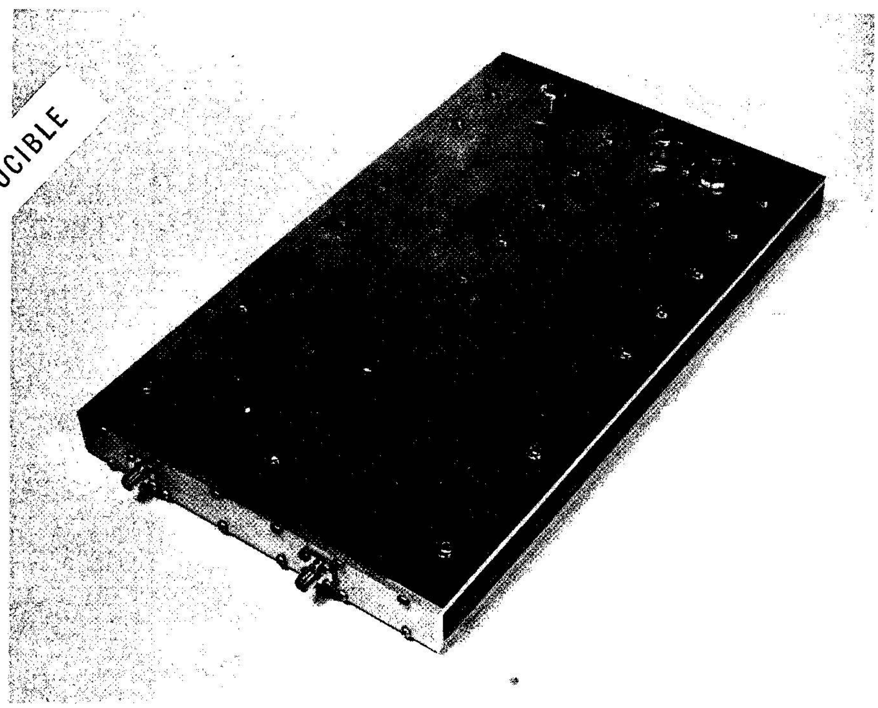


FIGURE 3-22. THEORETICAL VSB FILTER RESPONSE WITH ONE SINGLE TUNED AND ONE DOUBLE TUNED TERMINATION



(a)



(b)

Figure 3-23. Vestigial Sideband Filter

Thus, for a continuing program, the design should be modified to include a larger, silver plated, center conductor with double-tuned circuits for both terminations. This arrangement should easily meet, and probably exceed, the response requirements of Figure 3-20.

3.5 MONITOR AND PROTECTIVE CIRCUITRY - TASK 5

3.5.1 Requirements of Circuitry

Performance monitoring is required for those devices and components of the transmitter where an electrical failure (due to shorts, voltage breakdown, etc.) can result in a catastrophic system failure if unchecked. Signals from monitoring devices are utilized to actuate protection and control circuitry to prevent permanent damage to critical components like the final power amplifiers and the power conditioner subsystem. Proper and instantaneous action when a fault occurs usually permits the circuit to be operated again after the fault is cleared or eliminated. Note that laboratory equipments used in this program do not include the features that would be necessary for adequately protecting the components of the breadboard transmitter.

Specific circuitry includes an electronic crowbar with appropriate sensors for dc protection, an rf drive switch with sensors for rf fault protections,and logic circuitry to effect protective action when a fault occurs. A dc breakdown, which would endanger the final amplifier stage if within a tube or the power conditioner if external to the tubes, causes the crowbar spark gap to be triggered, shorting the power supply output to ground in less than a few microseconds. At the same time, the logic circuit opens the prime power input bus to the power conditioner. RF drive is removed in the case of rf breakdowns in the waveguide assembly. After a suitable time delay, the logic circuit can turn the equipment on again, depending on the specific nature of the fault. Parameters monitored include plate voltages and currents, grid currents, RF input and output signals in both the forward and reverse directions, stage efficiencies, and gain and phase distortions.

Note that a small amount of arcover energy (< 10 joules) is usually beneficial to a tube since it will clear a flaw caused by a metal whisker inside the tube. During shutdown when an arc appears, the crowbar device must divert stored power supply energy and must remain a virtual short circuit across the load for a short time. If

the crowbar action were to cease before the main power source has been disconnected and storage devices thoroughly discharged, a voltage sufficient to cause a recurrence of the arc could easily built up again in the storage capacitor. The total energy stored in the Doherty power supply may be as much as 140 joules (2500 volts on a 45 μf capacitor) when the full power rating of 5 kW sync peak power is used.

The VSWR protective circuit uses a threshold detector operating on a rectified sample of the reflected RF power. When the threshold is exceeded (preset at a nominal maximum safe level) by a large power reflection, a signal is generated to remove RF drive and high voltage from the tubes.

3.5.2 Circuitry

The circuitry necessary to satisfy transmitter protective requirements includes monitor circuitry with necessary sensors and protective circuitry which utilizes monitor circuit outputs as triggers when faults occur.

3.5.2-1 Monitoring Circuits

Requirements for monitor signals to control grounded grid triode rf amplifiers are implied in the following emergency shutdown sequence when a fault occurs:

1. VSWR Trip shuts off RF drive (and also HV if desired) when RF reverse power exceeds a pre-determined level.
2. DC faults which turn off HV and RF drive are:
   - Plate Current Overload
   - Grid Current Overload
   - Crowbar Firing

   The HV power supply should be crowbarred when excessive current in any of the tubes is sensed. An auxiliary crowbar interlock closure is required to insure that the HV power supply is then turned OFF.

Figures 3-24 and 3-25 show the monitor locations and sensors for the three high power stages; the Doherty, driver, and aural. The following lists indicate specific monitoring points:

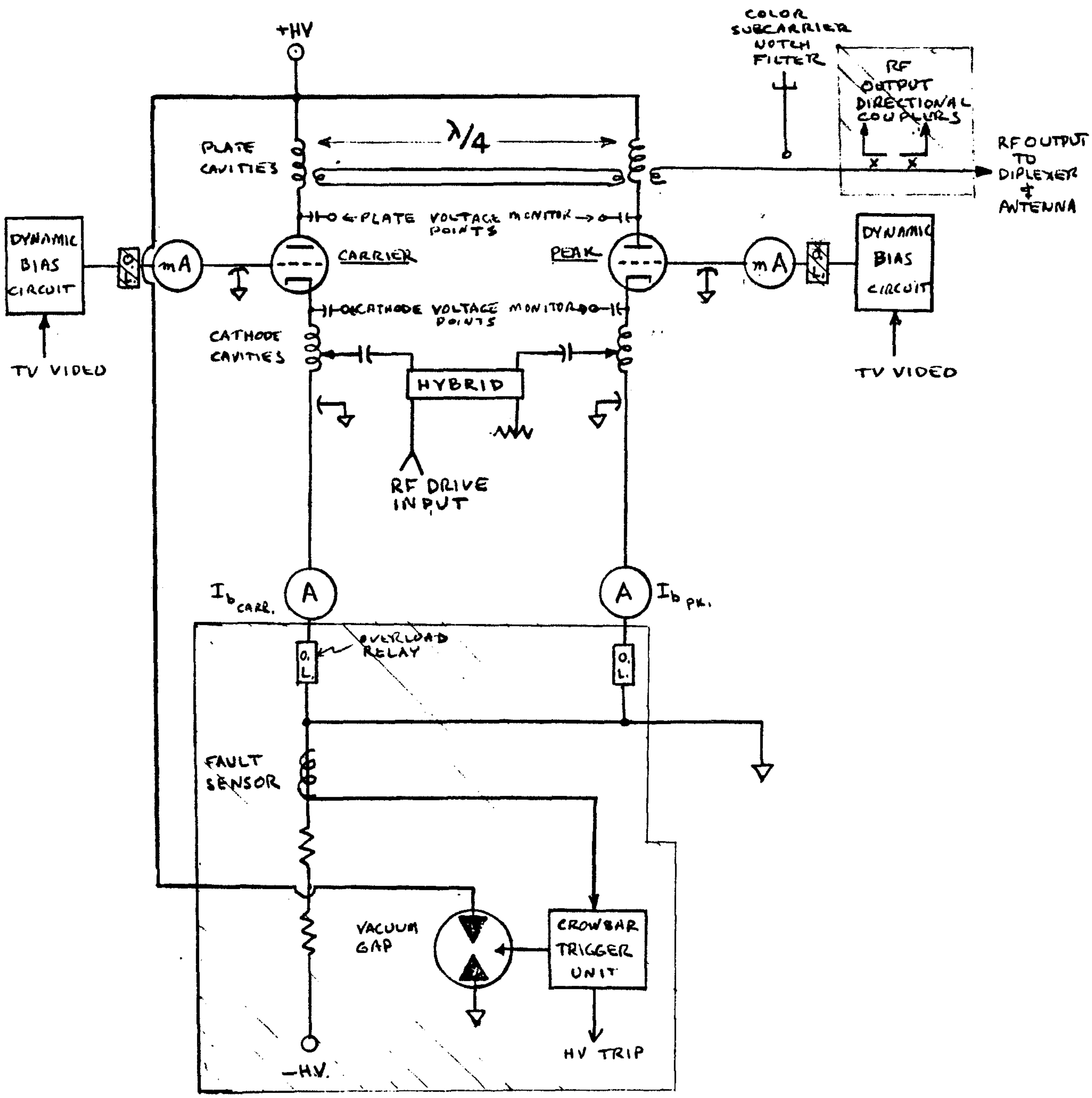


FIGURE 3-24. DOHERTY MONITORING AND PROTECTIVE CIRCUIT

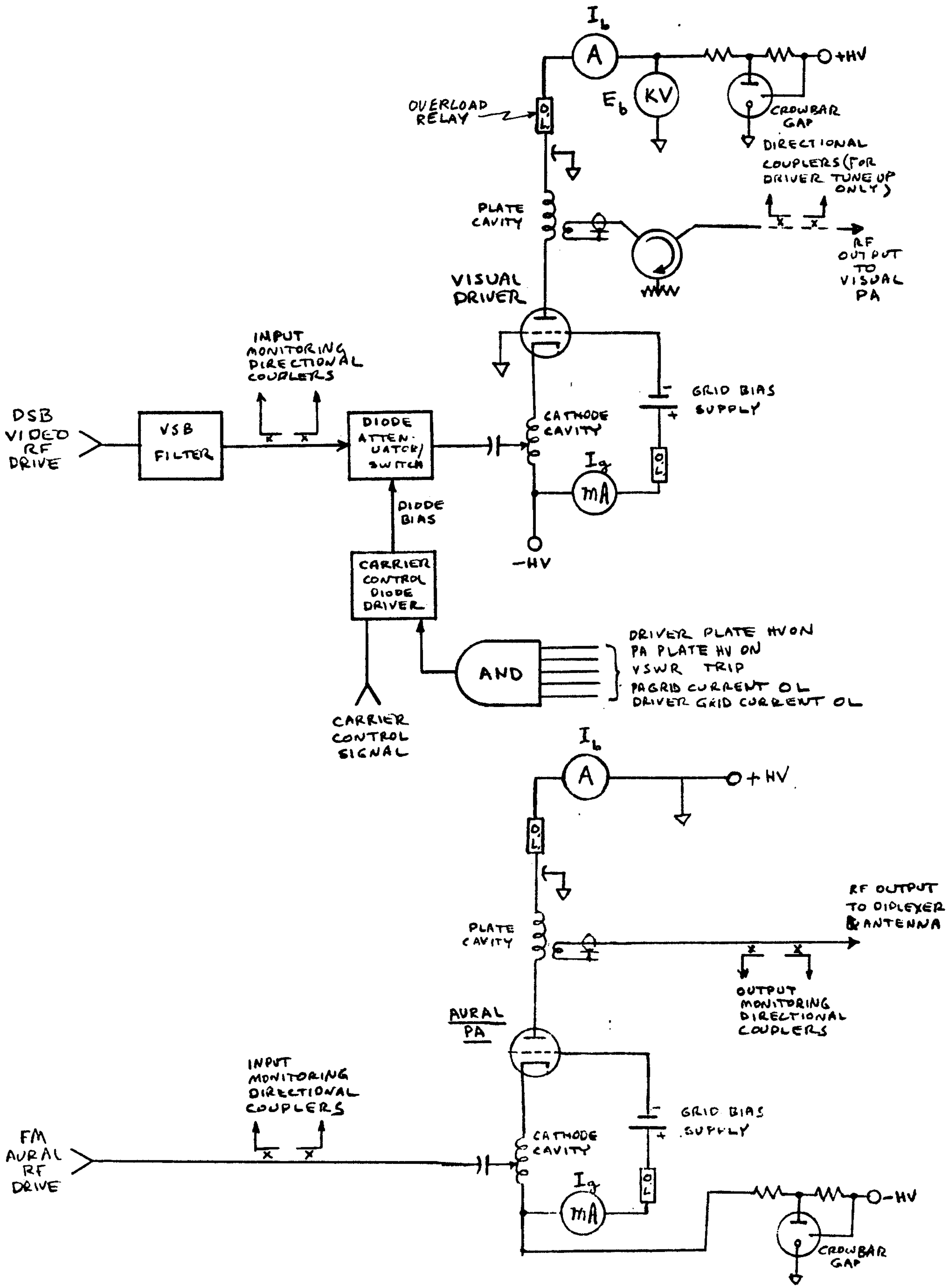


FIGURE 3-25. AURAL AND DRIVER MONITORING AND PROTECTIVE CIRCUITS

Doherty Visual Amplifier - Figure 3-24

- Carrier Tube Plate Current
- Peak Tube Plate Current
- Plate Voltage (common to both tubes)
- Carrier Tube Grid Current
- Peak Tube Grid Current
- RF Monitor Points
  - Carrier Plate RF Voltage Sample
  - Peak Plate RF Voltage Sample
  - Carrier Cathode RF Voltage Sample
  - Peak Cathode RF Voltage Sample
- Total RF Input Power (Forward and Reverse)
- Total RF Output Power (Forward and Reverse)

Visual Driver - Figure 3-25

- Plate Current
- Plate Voltage
- Grid Current
- RF Output Power (same sensor as Doherty input is suitable)
- RF Input Power (Forward and Reverse)

Aural Amplifier - Same as Visual Driver

3.5.2-2 Protective Circuits

The selection of the crowbar spark gap considered size, weight, ruggedness and reliability as well as performance. All circuitry is solid state; the design of the crowbar and associated circuitry was based on a need to operate under varied power supply and transmitter conditions. A vacuum spark gap was selected initially because of its wide range and ability to operate effectively at lower test voltages as well as higher operation voltages. This approach was later abandoned due to poor operating reliability and a gas spark gap tube used. The vacuum spark gap appeared to have some long-life limitations which were not evaluated specifically in this program. The trigger, logic, and control circuitry developed were solid state, using a commercial unit (TM-11) for the 30 kV trigger to fire the spark gap.

The crowbar circuit, Figure 3-26, provided fault protection entirely adequate for high power direct broadcast satellite transmitters. This unit performed quite

satisfactorily in discharging 72 joules (9 µf @ 4kV) of stored energy during preliminary testing. The transmitter breadboard will operate well with this crowbar, although it will have a 2.5 kV across up to 45 µf, or up to 140 joules of energy. The delay time between initiation of the fault and the firing of the crowbar is less than 1.0 µsec and could be reduced further if necessary. The total discharge energy in the protected circuit was calculated to be well below the 10 joule value estimated to be detrimental to tube operation. Details on the design and operation are included in Section 5.5.

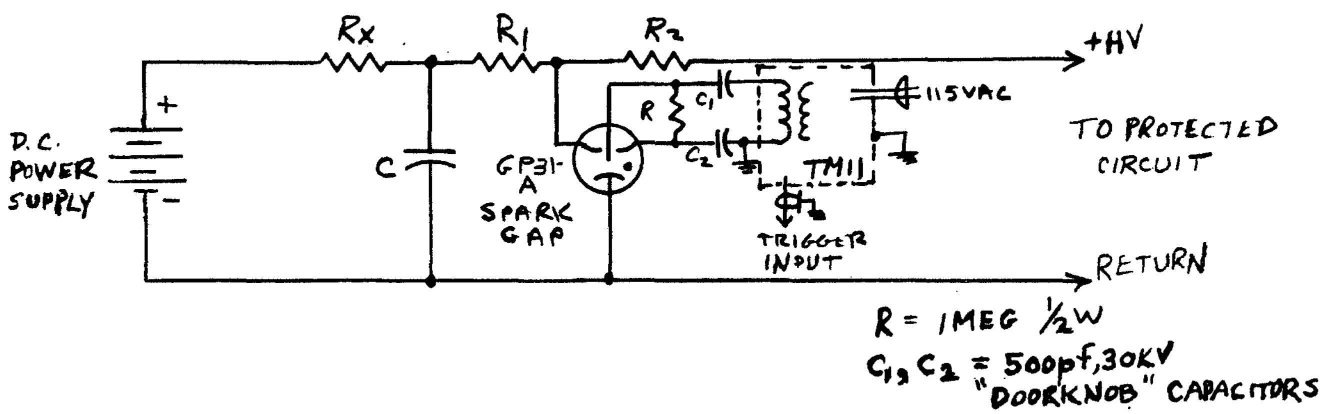


FIGURE 3-26. CROWBAR CIRCUIT

Two vacuum spark gaps were tested. The EG&G GP12BV required a large amount of stored energy at a high voltage to perform well, but was somewhat unreliable at low voltages and energies. A smaller GP20AV gap was then tried but again with somewhat disappointing results. The vendor indicated that previous experience had shown some erratic performance with these gaps. Finally, a GP31A gas spark gap was installed. This gap, however, which is excellent at the voltage and energy levels of the Doherty amplifier, does not work well below 1700 volts. Consequently, testing must be performed

with this limitation taken into consideration.

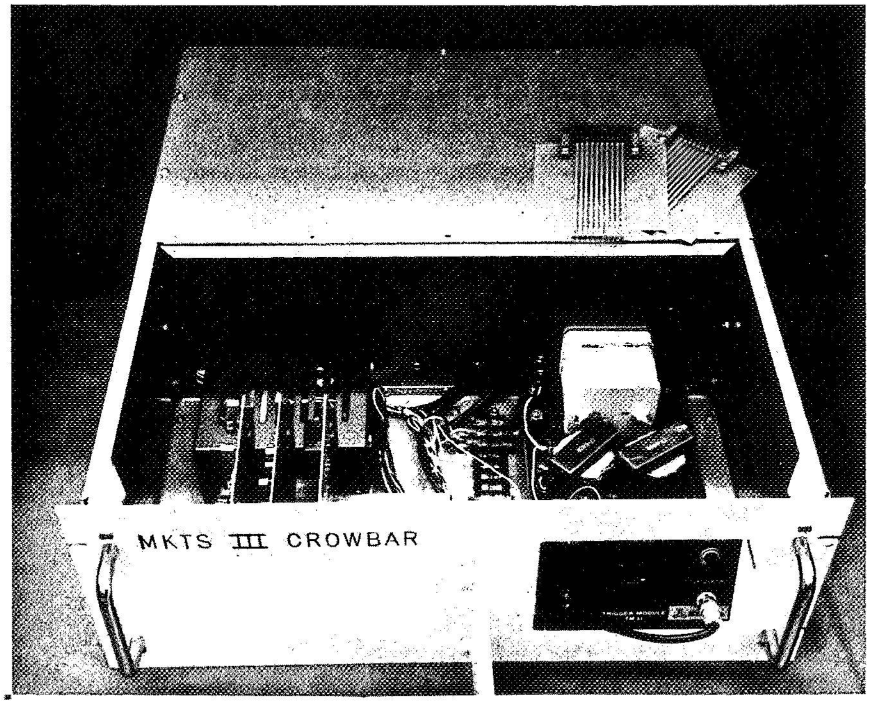
The VSWR trip circuit used, which is discussed in Section 5.5.4, will operate well over a broad range of RF input levels thus allowing maximum flexibility in the further design of RF circuitry for monitoring reflected power.

The complete unit is shown in Figure 3-27; the control circuitry is shown in (a), and the high voltage spark gap in (b). The entire unit is contained in an alodined aluminum enclosure measuring 18" x W x 20" D x 7" high; there is an access door to change trigger voltage on the TM-11 trigger module. The high voltage section contains the spark gap, two global resistors, the trigger isolation capacitors, and a one megohm resistor. The control part contains a modular ± 15 VDC power supply, the TM-11 Trigger Module with the ceramic pulse transformer output terminals feeding through to the high voltage section, and the trigger, control and logic circuits which are mounted on 4" x 4" printed circuit boards. Also included is the VSWR trip circuit on a separate printed circuit board.

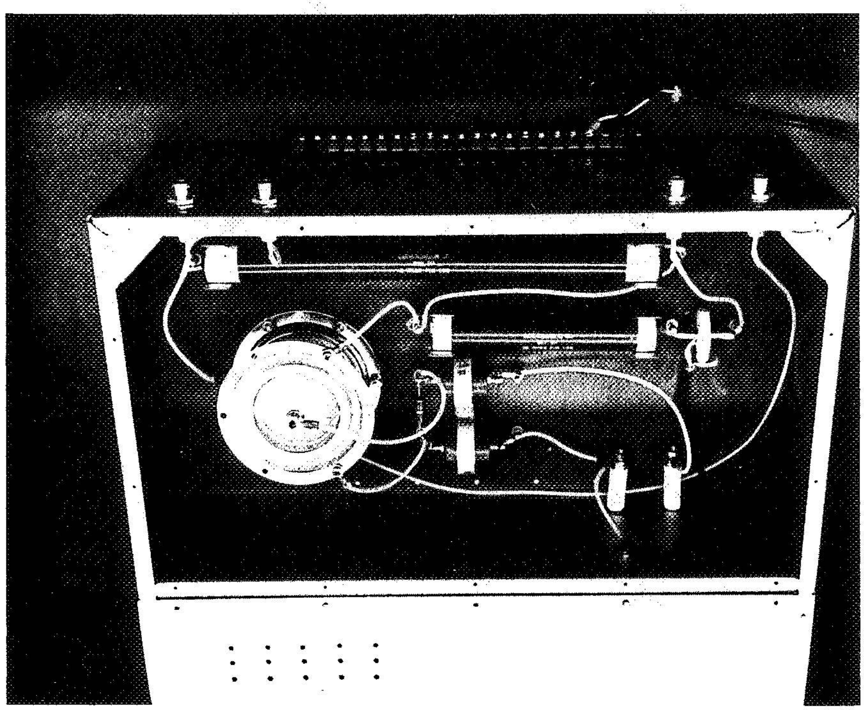
The crowbar for the high power visual amplifier stage uses an EG&G type GP31A vacuum spark gap. A separate crowbar circuit which protects the driver stage uses the EG&G KN-2 Krytron spark gap with an EG&G TR149 trigger transformer. The KN-2 is a cold cathode switch tube with a holding anode configuration which has a particle emitter to speed ionization. All of the system power supplies have a rapid turnoff device which operates when the spark gap is triggered, thus turning off the high voltages.

3.5.3 Test Results

A vivid qualitative demonstration of crowbar circuit performance is the foil test. A piece of 0.5 mil (.0005") aluminum foil was placed at the high voltage potential. The power supply charged the 9 μf capacitor to 4 kV; the ground lead was then brought into close proximity and an arc started. The results, shown in Figure 3-28a, show only a small arc track on the foil; the same test produced holes of approximately

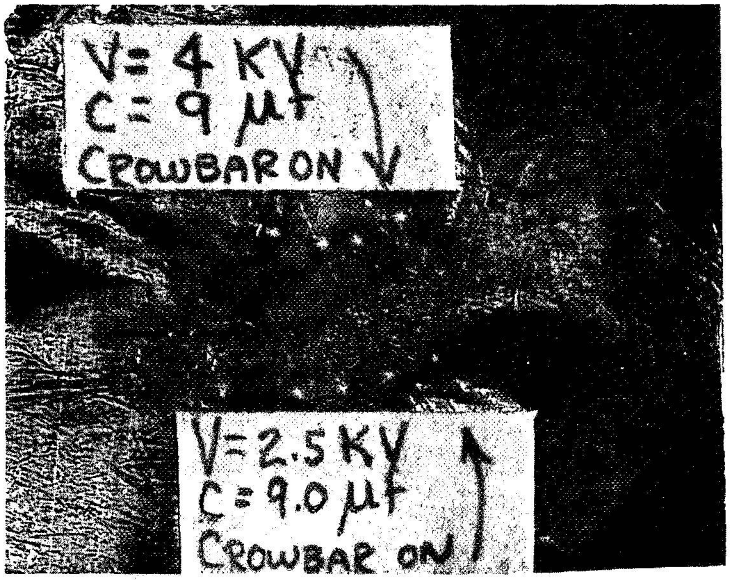


(a) Control Circuitry



(b) Spark Gap Circuitry

Figure 3-27. Crowbar and Supplementary Circuitry



a) With Crowbar

b) Without Crowbar

Figure 3-28. Results of Crowbar Tests

1/8" diameter when performed without the benefit of crowbar protection, as in Figure 3-28b.

The driver stage crowbar using the EG&G KN-2 Krytron was tested using the foil test. With the voltage set at 1.0 kV, the nominal supply voltage, no visible damage to the foil resulted, indicating that the crowbar is sufficient to provide full protection of the driver stage.

3.6 CONTROLLED CARRIER CIRCUIT DESIGN - TASK 6

3.6.1 Approach

This task was to design a "controlled carrier" modulator, or attenuator, for use with the 5 kW AM-TV transmitter breadboard. This circuit will permit the power supply and conditioner to be sized to the "average" transmitter power required for the TV signal. Carrier (or envelope) reduction for dark pictures has the effect of reducing the effective S/N at the receiver by 1 or 2 dB, which would normally not be noticed by the viewer. However, the technique can reduce power supply and conditioner capacity requirements by as much as 40%, which is highly significant to the satellite and system.

The approach in this task was to develop a circuit from the functional block diagram of Figure 3-29, using the performance specifications of Appendix A as a design guide. The resulting circuit is arranged to permit ease of parameter changes during testing although an operational design, based on this circuit,would not require as much flexibility.

3.6.2 Circuit Design

The circuit is based on the stages shown in the block diagram, Figure 3-29 and includes the functions of the power sampler and the four blocks in the left half of that figure. The final schematic diagram, shown in Figure 3-30 includes the power sampling circuit using a 2 ohm resistor, R1, in series with the B-minus line a threshold control in conjunction with the time-constant control circuit, the differential amplifier (using

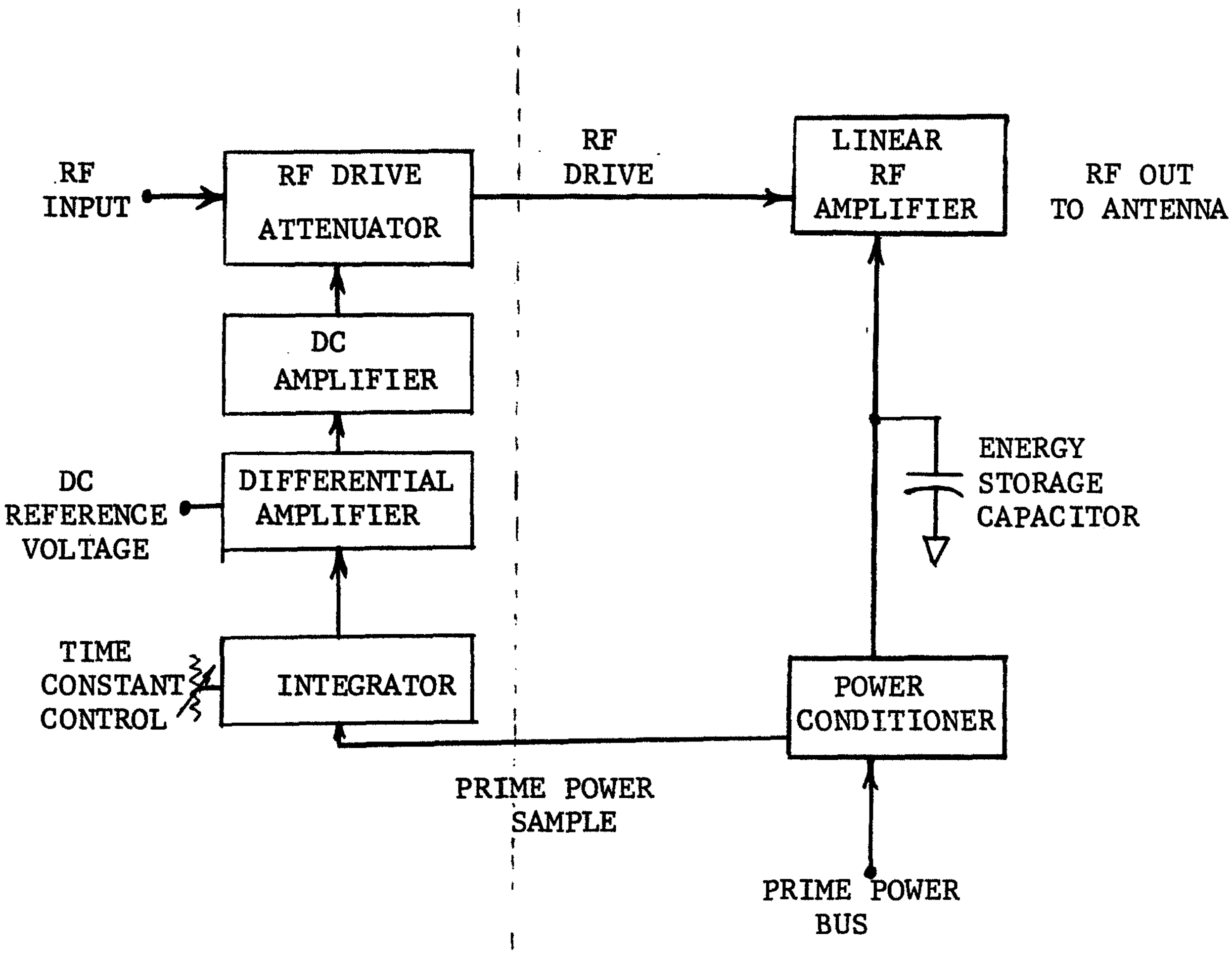


FIGURE 3-29. BLOCK DIAGRAM OF CONTROLLED CARRIER CIRCUIT

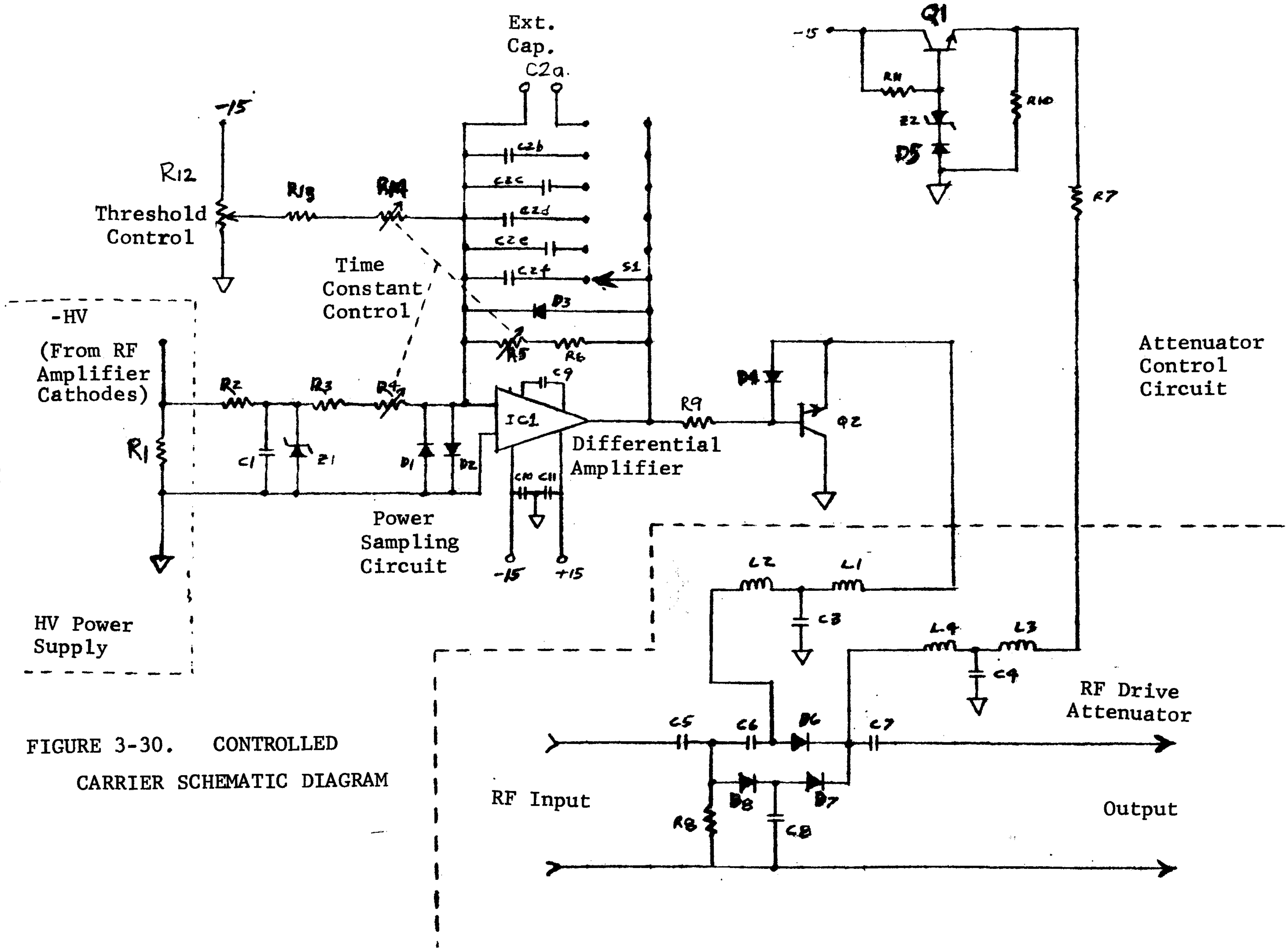


FIGURE 3-30. CONTROLLED CARRIER SCHEMATIC DIAGRAM

IC-1), and the RF drive attenuator with its attendant attenuator control circuit. A detailed description of circuit performance is given in Section 5.6.

The circuit varies the signal level of the driver input by up to 6 dB, although this range is not required for most TV signals. The unknowns in signal pre-emphasis makes a safety margin desirable, however. Computed operation is;

Insertion loss = .282 dB
Range = 6 dB

Testing with this circuit was deferred until the high powered Y2042 tubes were available for testing in the transmitter.

3.6.3 Fabrication

The circuit was fabricated using strip line techniques for the RF Drive Attenuator section; the Control Circuit was fabricated on a vectorboard base. Figure 3-31 is a photograph of the completed unit.

3.6.4 L-C Filter

An integral part of the Controlled Carrier concept is the relation of the L-C energy storage filter to the control circuit response. A large L-C filter would reduce the burden on the Controlled Carrier function and would be used if weight and cost were no limitation and the spark gap protective circuit could handle the stored energy. Practically, a small filter is desirable with the Controlled Carrier circuit providing the necessary compensation. From previous experiments[7], a time constant of two to six milliseconds was found to be acceptable for conventional TV programming material. For the purpose of this program, a 40 μf filter capacitor was selected. The effectiveness of this selection would be basically determinable by TV viewer reactions and should be evaluated to a greater extent after a specific mission is identified and design of an operational system implemented.

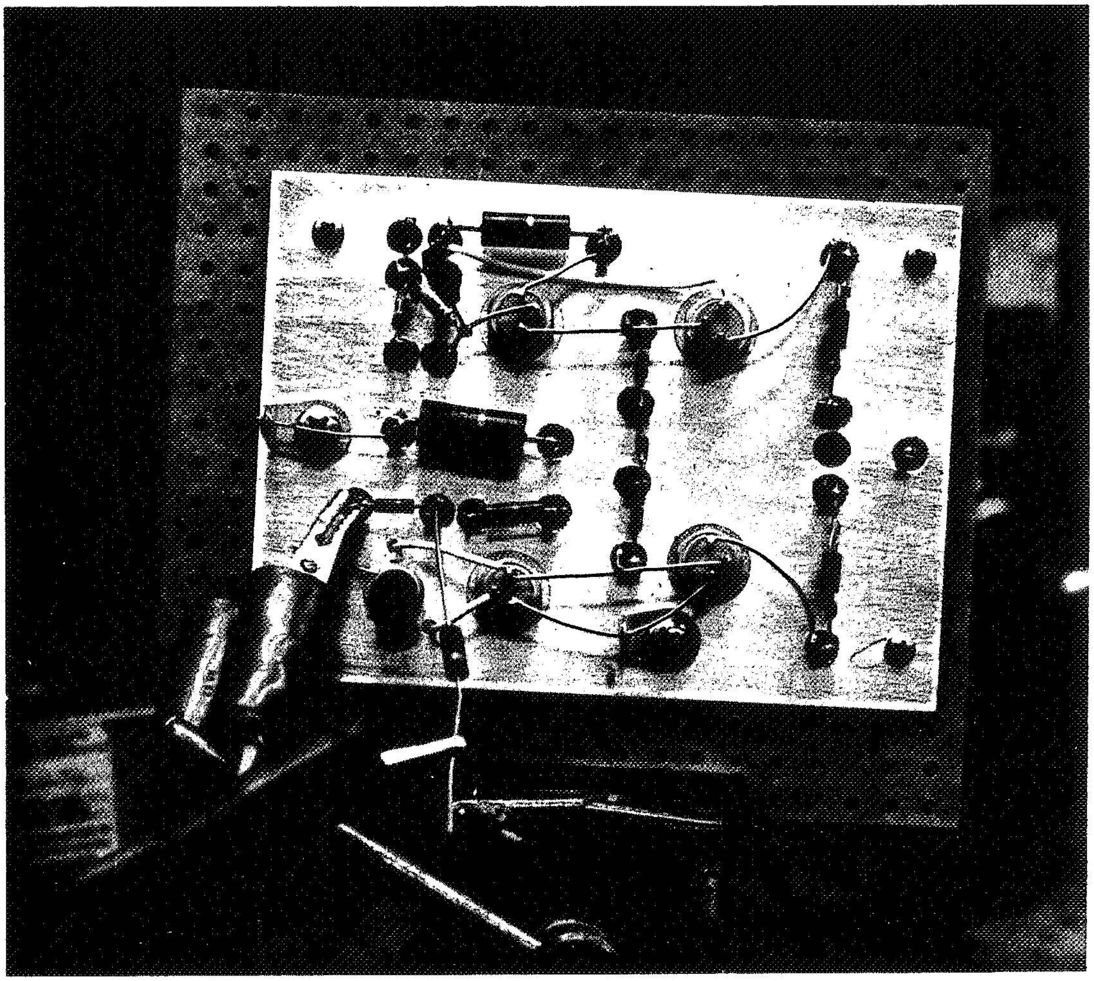


FIGURE 3-31. CONTROLLED CARRIER EQUIPMENT

3.7 HIGH POWER RF COMPONENT ENVIRONMENTAL TESTING - TASK 7

3.7.1 Test Philosophy

A basic test program was formulated for multipactor and ionization breakdown tests in a simulated space environment on selected components which are representative of ones which may be used in a high power satellite transmitter. Tests were to be performed at power levels up to 2.5 kW in a vacuum chamber; scaling of data can then permit estimates of performance at higher power levels. High power operating and breakdown relationships were considered extensively in a previous contract.(2)

The test plan detailed in Section 5.7 describes objectives, components, techniques, test methods, and limits imposed in testing RF components for high power multipactor and ionizing breakdown under high vacuum conditions. Also, evaluation of means to avoid breakdown through suppression techniques and component design configurations were considered. Test results were to provide vital experience and data on the occurrence and suppression of multipactor and ionization breakdown, identifying critical parameters and corrective measures to suppress breakdown. Guidelines for the selection of rf components for other systems are a part of the results expected from these tests. Components selected for testing were:

- 3-1/8" coaxial line
- half height WR975 waveguide (2.5 x 10 inches)
- Stepped 3-1/8" coaxial line
- Stepped WR975 waveguide

Circumstances permitting, the 3-dB hybrid and color notch filter would also be tested although neither were expected to show any multipacting tendencies.

3.7.2 Results

A roadblock in achieving the stated objectives was the inability of the rf source to provide the required output power. The Y1498 tube, used in the aural stage while awiating the Y2042 type, was capable of only about 500 watts CW. The attempt was made to operate at 2.5 kW with 20% duty cycle, but tube gain was 7 to 10 dB less than

the Y2042 specification, and the required drive power could not be obtained with the test facilities originally considered adequate. Further testing was thus deferred until either the Y2042 tube was received (now considered for next year) or the Doherty amplifier with its high powered driver could be made available. The amplifier also was not available within the program period, however.

The 3-1/8" coaxial line and the half-height waveguide were tested at up to 700 watts; no breakdowns were expected and none occurred. Stepped sections of coaxial line and waveguide were prepared and are available for future testing. These stepped sections are designed to multipact below 2.5 kW and should show definite breakdowns within the capability of the Doherty/driver circuitry. After calibrating waveguide breakdown conditions, schemes to reduce susceptibility would be implemented.

In all the tests, instrumentation for monitoring vacuum level, temperatures, multipactor action, and ionizing breakdown were incorporated into the test setup. Figure 3-32 shows the basic test assembly used.

3.7.3 Test Parameters

The basic operating parameters for multipactor testing of the RF components were as follows:

1.	Frequency	700 to 900 MHz
2.	RF Power	to 2.5 kW CW or Peak
3.	Temperature	500° F max.
4.	Pressure	Approx. 10^{-6} mm Hg

During test experimentation, other relevant parameters to be measured included:

a) VSWR
b) Incident and Reflected Powers
c) Reflection Coefficient
d) Breakdown Current (Multipactor Electron Density, if any)
e) Breakdown Voltage
f) Gap Spacing
g) fd product (MHz-cm)
h) Materials (including surface treatment or other processing)
i) RF Pulse Shape if not CW

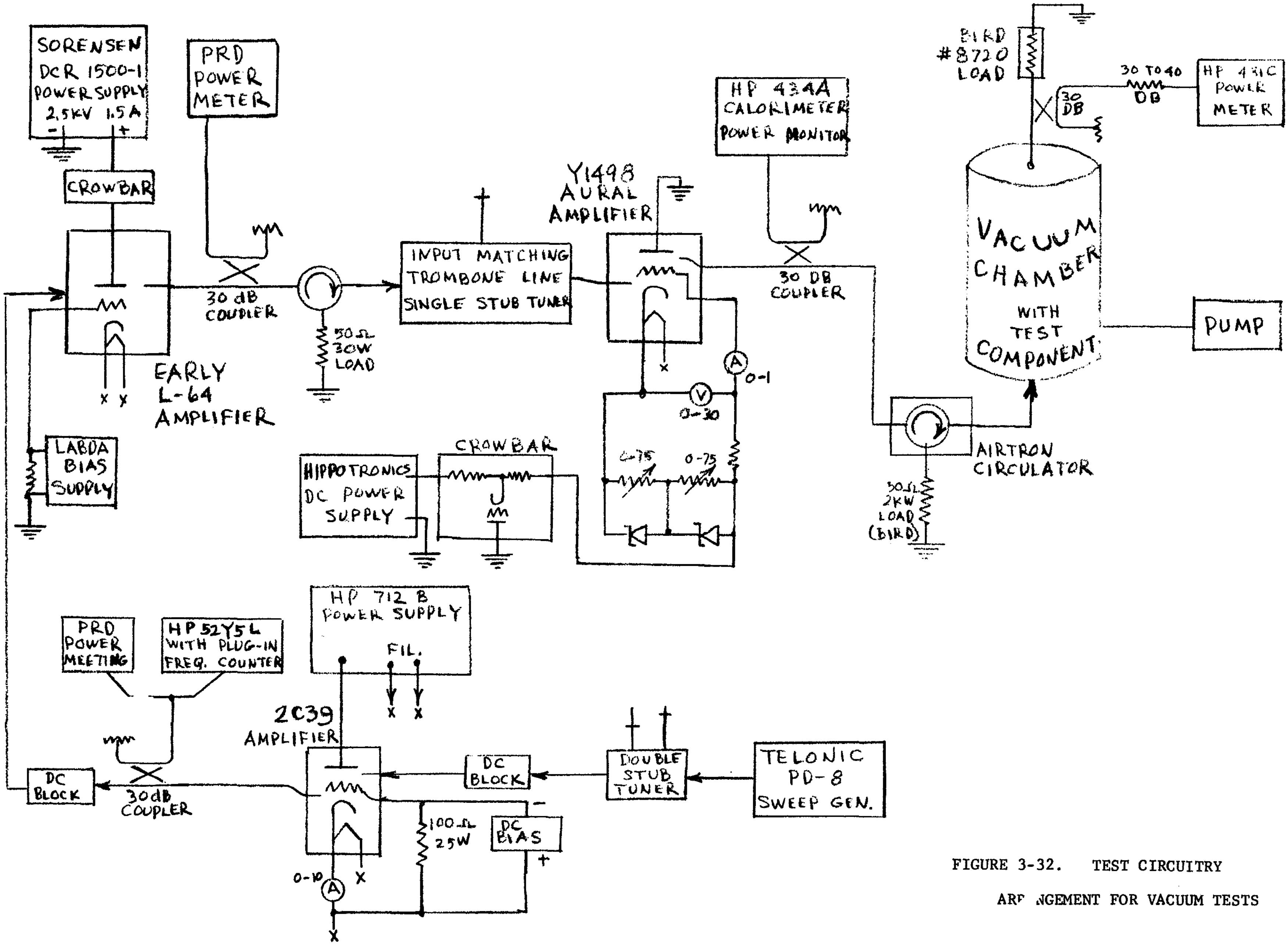


FIGURE 3-32. TEST CIRCUITRY ARF NGEMENT FOR VACUUM TESTS

3.7.4 Components Descriptions

The first tests were concerned with obtaining satisfactory operation of the vacuum test system with uniform coaxial line and waveguide components. After this has been accomplished, a waveguide component and a coaxial component with stepped gaps, shown in Figure 3-33 b and d, designed deliberately to multipact, would be tested. These components will permit debugging and calibration of sensors for multipacting and other breakdown phenomena measurement. The test components are also to be used in the evaluation of multipactor suppression methods. Items (b) and (d) in Figure 3-33, the stepped waveguide and coax sections, can be considered as presentative elements of reactive harmonic filters that might be used in a high power coaxial or waveguide transmission system, and are also representative of low impedance lines (coaxial, rectangular waveguide, and ridged waveguide) which are a possible approach to multipactor prevention. Both stepped units were designed to accommodate variations in gap spacing, disassembly, inspection and cleaning. The stepped section components can also be used in the proposed test of flame sprayed materials which prior investigations found to be useful in multipactor suppression.

3.8 TRANSMITTER TESTING - TASK 8

3.8.1 Test Requirements

The transmitter test plan was designed to provide direction for obtaining performance data on the transmitter, including the measure of its capability for performing as a TV transmitter. For an AM-TV space mission, the Controlled Carrier feature (Section 3.6) is considered necessary, and system tests as a TV transmitter should include that circuit. The specific parameters and performance characteristics included in the test plan are shown in Table 3-1; the list includes the fundamental operating measurements, fundamental TV measurements, and the various specialized tests and refinements to the test program. Separate aural channel measurement requirements are also indicated in the table. A list of the required test equipment is in Appendix C, and the methods of measuring each of the items in Table 3-1 are pre-

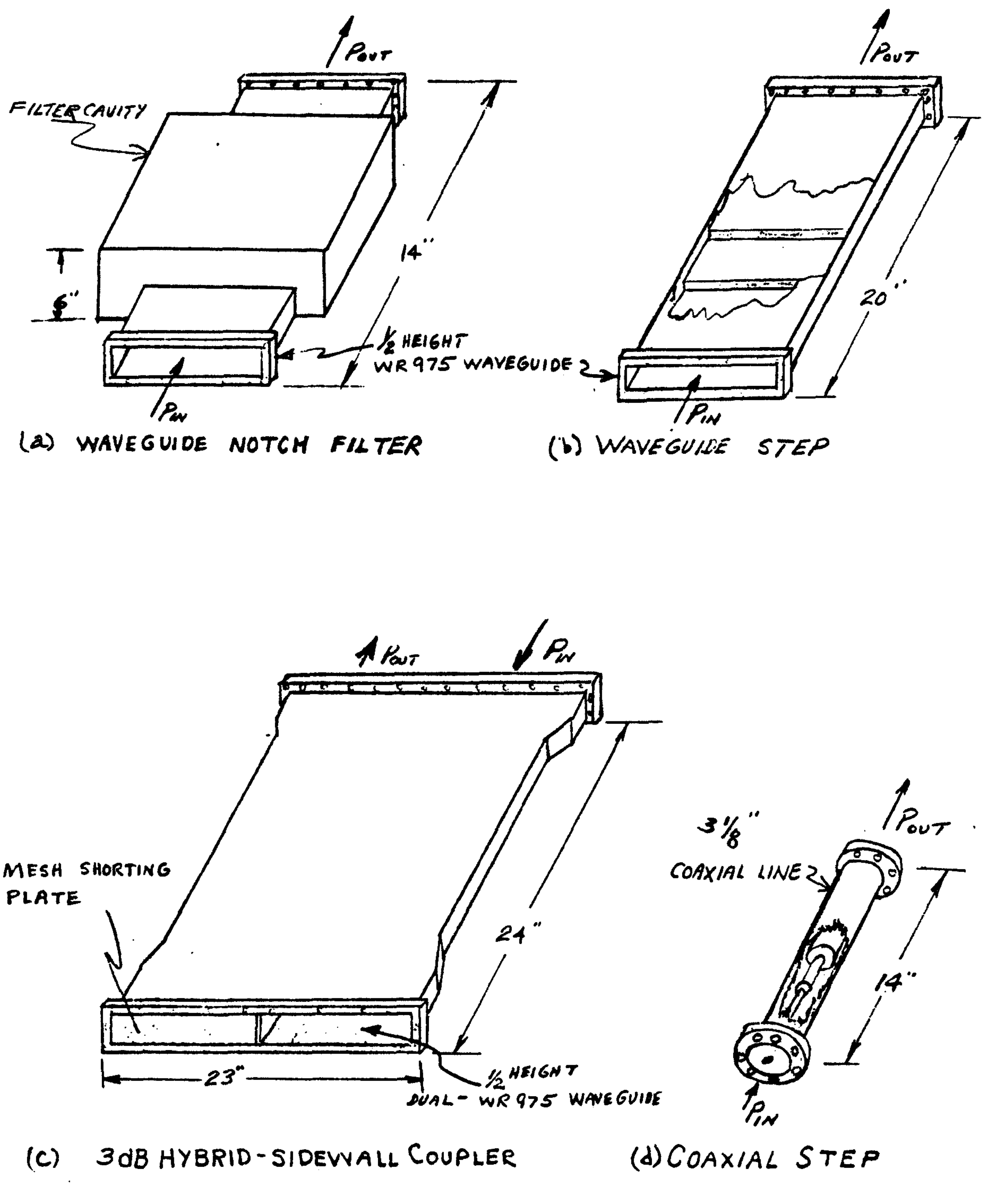


FIGURE 3-33. REPRESENTATIVE COMPONENTS SELECTED FOR MULTIPACTOR TESTS

TABLE 3-1

MULTIKILOWATT TRANSMITTER PERFORMANCE TESTS

1. Efficiency as a function of RF drive level and TV picture content

   (a) Operating voltages and currents
   (b) RF power outputs
   (c) Power dissipation factors
   (d) Power gains

2. TV picture quality factors

   (a) Frequency response
   (b) Linearity (low frequency)
   (c) Differential gain
   (d) Differential phase
   (e) Envelope delay
   (f) Hum and noise

3. Harmonic and spurious outputs

4. Controlled carrier operation

5. Power supply regulation and effects of transient loading (i.e., during vertical sync interval) with and without controlled carrier operation

6. Upper and lower sideband attenuation in visual RF channel with VSB and color image filters

7. Aural channel modulation performance

   (a) Operating parameters
   (b) Transmitter bandwidth
   (c) Transmitter contributed AM

sented in Section 5.8. Test procedures follow the EIA suggested methods whenever applicable. Simultaneous operation capability of aural and visual channels are used where appropriate.

The basic test exciter unit, assembled from items 3 and 12 through 17 in Appendix C, is shown in Figure 3-34. The exciter unit is designed to provide all the signals necessary for TV performance tests as well as the other parameter testing.

3.8.2 Test Facility

An overall testing facility was assembled to provide data on the complete transmitter, based on the test requirements above and the basic test exciter diagram of Figure 3-34. The overall instrumentation for the transmitter tests is shown in Figure 3-35.

3.8.3 Test Results

The major transmitter test results are included in specific component task results discussed previously. The Doherty amplifier and its driver were tested with water cooling to simulate the heat pipe operation that would be required in space. In addition, the transmitter was mounted on a cooled base plate. Crowbar protective circuits were used for the three high power circuits (aural, Doherty, and driver). Results of tests on these three stages are essentially the same as the individual stage results in Sections 3.2 and 3.3. Results in the TV tests performed did not differ when the circuits were combined into a transmitter since interfaces are basically linear in character. The aural stage is essentially independent of the visual stages and met all requirements except that the Y1498 tube caused a minor gain limitation. The Doherty amplifier was tested after first measuring the performance of the carrier stage as an independent Class B linear stage. As noted previously, efficiency increased by as much as 58% for the cases considered, which is about the best to be expected. Lower efficiency differences at smaller signal operating conditions were expected and observed.

These test results indicate the superiority of the Doherty over the linear Class B

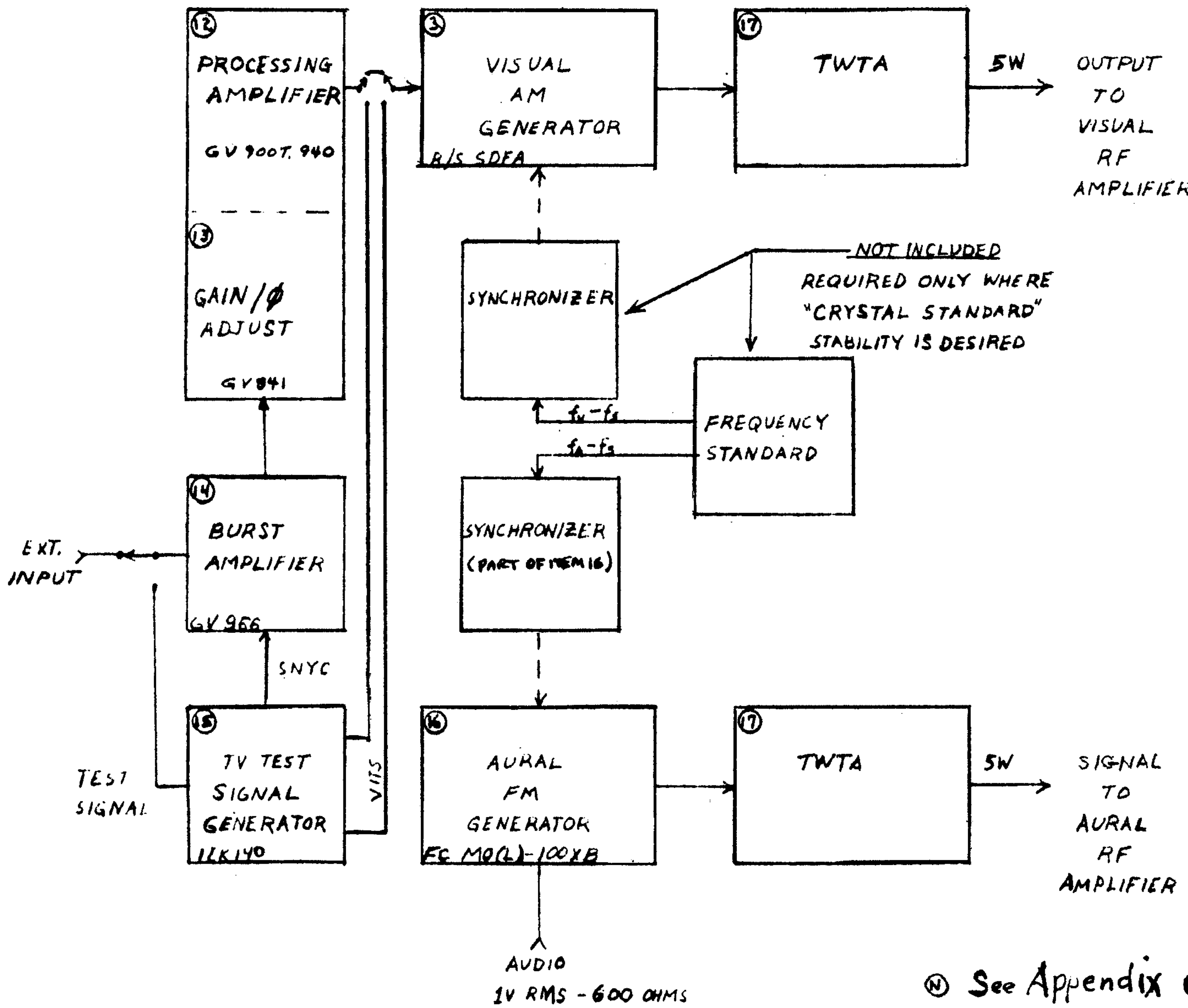


FIGURE 3-34. TEST EXCITER BLOCK DIAGRAM

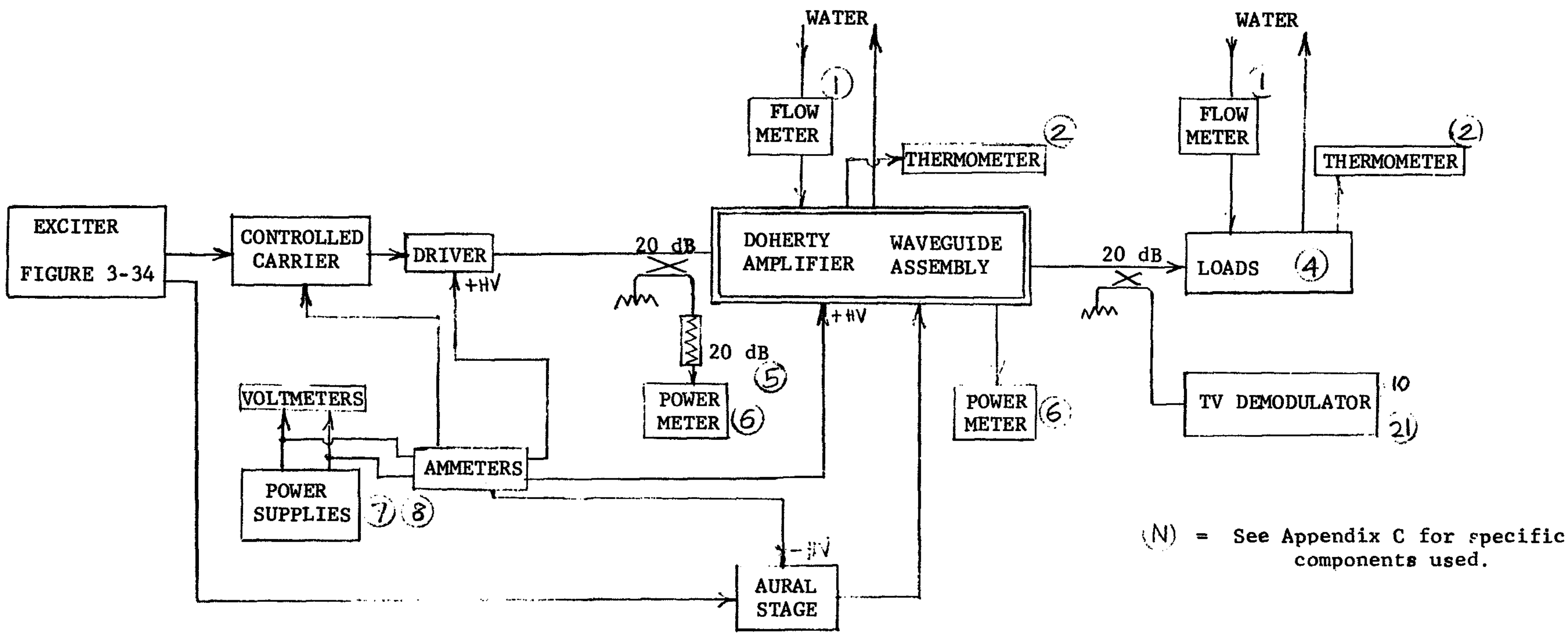


FIGURE 3-35. SYSTEM TEST INSTRUMENTATION

stage. However, further tests with the Y2042 tubes, when obtainable, should be implemented to obtain higher power tests which should be much more meaningful.

3.8.4 Test Descriptions

A brief description of the purpose of each of the tests provides a basis for formulating the overall breadboard transmitter tests, outlined in Section 5.8. The brief descriptions which follow are based on the list of Table 3-1.

3.8.4-1 Efficiency Tests

The first set of tests measured transmitter operating parameters as a function of TV picture statistics, using a stair-step waveform to simulate the TV picture. These tests are designed to provide a measured baseline of operating parameters (without the controlled carrier function in operation). Results were read directly from the efficiency meter included in the test facility.

3.8.4-2 TV Picture Quality

The frequency response test in group 2 of Table 3-1 is intended to establish the overall video amplitude versus frequency response of the transmitter and to verify that the results are in accordance with the EIA standards. The low frequency linearity test is used to establish the output amplitude versus input amplitude relations for the transmitter. This test also indicated the compliance of the transmitter with paragraph B-9 of EIA Standard RS-240 (or equivalent) for visual broadcast equipment. The test measured the differential gain of a 3.58 MHz signal (color subcarrier) as the average picture level (APL) varied in the stair-step fashion mentioned above, showing very little variation from a linear characteristic.

The differential phase test measures the differential phase of the signal under the same varying picture levels (paragraphs B-11 and B-10 of RS-240). Envelope (or group) delay versus frequency for the visual channel of the transmitter and hum and noise tests were not performed specifically, but should be done in accordance with

the EIA Standard when the transmitter is operated at full power.

3.8.4-3 Harmonics

The harmonic output test is designed to measure harmonic, subharmonic, and spurious radiations from the transmitter as an indicate of the degree of filtering required in future transmitters of similar design in order to insure that these radiations are at least 60 dB down from the peak visual carrier power level. Quantitative data was not considered significant until the higher powered Y2042 tubes should be obtained.

3.8.4-4 Controlled Carrier Tests

The controlled carrier tests permit evaluation of the controlled carrier mode of transmitter operation. The measurements for these tests should be compared with those taken for the transmitter in the normal mode of operation, and judgements made as to the relative effectiveness of the controlled carrier mode. Again, this operation was deferred until higher power is used.

3.8.4-5 Other Tests

The power supply regulation test is designed to evaluate the performance of the power conditioner unit and LC filter when subjected to normal transient conditions. These transient conditions might be caused by radical changes in the picture content of the visual carrier. This test, performed without the controlled carrier mode, showed adequate L-C filtering. However, an evaluation of the controlled carrier mode is desirable before forming final conclusions.

3.8.4-6 Aural channel tests

The last group of tests is concerned with the aural channel of the transmitter. The operating parameters tests measured the input power requirements, rf output power, and quantity and mode of heat dissipation for this section of the transmitter. Efficiency from the measurements was somewhat less than computed. The aural channel bandwidth test was designed to measure the bandwidth characteristics (such as passband

flatness) for this channel; results indicated an acceptable aural channel characteristic. The final test to measure hum and noise modulation present in the rf output signal amplitude was deferred until a Y2042 type tube becomes available.

SECTION 4

RECOMMENDATIONS

4.1 GENERAL

The success in demonstrating the high efficiency capability of the Doherty linear amplifier relative to the conventional Class B linear type leads to the following overall recommendations:

- The Doherty amplifier is very effective in the UHF band for providing a high average efficiency for AM signals,
- The average AM efficiency substantially exceeds that of a Class B linear or klystron amplifier as normally used in TV transmitters,
- The transmitter breadboard is recommended for development into a complete and efficient transmitter system for an AM-TV broadcast satellite system. It is also recommended for any other system requiring good linearity with high efficiency.

In addition to the above general recommendations, each task resulted in a set of special recommendations for component improvements which provide a guide to the development direction for an eventual flyable equipment.

4.2 TASK RECOMMENDATIONS

4.2.1 Transmitter System Design - Task 1

- The transmitter performance should be based on the EIA RS-240 TV standard as is used here.
- The Y2042 tube (long-life production version of L-64S) is recommended for the output stages for high efficiency at high power; the driver can use an ML-8536 or equivalent. The Y1498 has been used because of Y2042 unavailability at the present time; the former uses a conventional cathode with a shorter inherent lifetime.
- Other recommendations of the system design task are included in the specific task recommendations that follow.

4.2.2 Visual Channel Amplifiers - Task 2

<u>Driver Stage</u>:

- A grounded grid stage with double-tuned output circuit is more than adequate to drive the Doherty amplifier; the ML-8534 provides a considerable power margin over requirements.

Doherty Amplifier:

- The Y1498 has provided adequate data to demonstrate high efficiency amplifier operation in the upper UHF-TV bands.

- The future Y2042 tube will be required to achieve the 5 kW peak power objective. Availability may be in mid-1971.

- The cavity designs for the Doherty amplifier provide the flexibility desired for tuning and matching adjustments, which are more complex than in a single stage amplifier.

- One likely form of input circuit is a 3-dB hybrid which procides both the required power split and the -90^{o} phase shift required between stages. A delay line type should be considered further to simplify biasing.

- Dynamic bias circuits can be included in both stages to obtain good linearity and a high efficiency. However, a further design effort may simplify the biasing, which is considered further in Section 5.2.2-2, and should be investigated further before a final engineering model is fabricated.

4.2.3 Aural Channel Amplifier - Task 3

- The grounded anode configuration encountered a problem with the grid bypass capacitor. To operate the amplifier as designed, include a choke section in the capacitor which then results in good input/output isolation.

- The anode breakdown problem from multipacting should be checked in a vacuum environment, although no serious problems are anticipated.

- Thermal control for a grounded grid circuit presents additional difficulties which were evaluated from a system viewpoint. A large capacitor is required to provide a low impedance thermal path.

- Initial testing was performed with loop output coupling, but iris coupling should be used in an operational amplifier.

4.2.4 RF Components - Task 4

High Power Components:

- All are based on half-height WR975 waveguide; coaxial line is susceptible to electrical breakdowns and thermal dissipation difficulties, and should be only used for very low powers.

- The rf components developed can be used directly; they meet or exceed specifications. An additional tuning means is necessary to obtain the 1.1 VSWR desired for the overall assembly.

- A 3 dB hybrid was used for testing, but it is not required for general tests. A complete diplexer with suitable isolation filters is required for an operational system. These have been discussed in earlier studies.[(2)]

Vestigial Sideband Filter:

- A low power filter fabricated for the test transmitter is preferred; it precedes the driver stage, and performance is considered adequate. Future transmitter tests should assess the effects of non-linearities in the high power stages to a greater extent than was possible here.

- The Q's of the reactive terminations were not adequate for the system requirements. The Q's may be increased by reducing the stripline loss for an operational system. This will require either silver plating and/or both terminations should probably be employed.

4.2.5 Monitor and Protective Circuitry - Task 5

- Monitor and protective circuitry must include crowbars for dc protection and suitable RF sensors to detect RF breakdowns; an RF control circuit is required to prevent rf faults from possibly disrupting or damaging the system.

- Logic circuitry developed for protection against dc arcing and high reflected rf power levels will turn off equipment to eliminate destructive effects of breakdowns. The equipment developed incorporates logic control circuitry in the final transmitter system.

- The vacuum spark gaps were judged less reliable than the gas spark gaps; the latter are recommended although their operating voltage range is restricted and system operation must be adapted accordingly.

4.2.6 Controlled Carrier Circuit Design - Task 6

- A controlled carrier circuit is necessary for a space AM-TV transmitter where conserving dc power is vital to achieve high system efficiency.

- A circuit suitable for the controlled carrier function has been designed. Its operation should be implemented at an early time in a program to assure adequate performance capability.

- A threshold circuit for actuating the controlled carrier circuit is used; setting a threshold arbitrarily can affect the amplifier efficiency and power supply size. This compromise should receive further attention. Initially, it appears best to set the threshold at the overall average power level but this may be a function of signal statistics and requires additional analysis.

4.2.7 High Power RF Component Environmental Tests - Task 7

- High power tests on rf components in a vacuum environment provide a way for assuring a space oriented transmitter system which will not be prone to rf electrical breakdowns.

- The program in this contract was to cover components applicable to a UHF TV transmitter system; results can be extrapolated to estimate operation of components at other frequencies.

- The Doherty amplifier, when equipped with the Y2042 triodes, would be a preferred test-signal source; the single Y1498 could not provide sufficient power.

4.2.8 Transmitter Tests - Task 8

- The method used for testing this transmitter, which checks for both functional performance (power, efficiency, stage gains, bandpass characteristics) and AM-TV performance based on the EIA Standard RS240A, is considered the best overall test approach for the type of transmitter represented by the breadboard.

- Results of the test suggest that the transmitter breadboard test assembly be left intact with the objective of retesting with the Y2042 higher power tubes when they become available. These tubes use a unique "bi-potential" cathode which reduces grid dissipation dramatically, and permits the high power operation to 5 kW peak to be achieved without destructive grid heating. The Y1498 tubes without the special cathode configuration as used in the tests in this program can safely achieve about 1 kW in a Doherty CW application.

4.3 FUTURE FOLLOW-ON RECOMMENDATIONS

A continuation of the program of this contract is largely dependent on programs and missions to be proposed in the near future. The techniques which evolved are ideal for any system requiring linear operation, which includes AM systems and multiple-carrier transmission with any type of modulation. For the latter type of system, the inherent linearity of triodes in amplifiers of the Doherty type permits a higher average efficiency than other amplifier types while retaining a very small intermodulation distortion level.

The proposed direction of effort would include further evaluation of tubes, including life testing in the UHF amplifier circuit when the Y2042 types are obtained. The transmitter evolution would continue, but be adapted to requirements of an established mission. The transmitter itself should be fabricated essentially as in the development of this program. The suggestions for changes include the following possibilities:

- the aural stage and the Doherty stages should all be dc grounded grid or grounded plate to simplify the power supply and protective subsystems;
- tuning adjustments may be modified in a final version since basic cavity design has been determind with the very flexible adjustments used for the breadboard Doherty;
- flight model development would follow the same basic methods used for any other satellite transmitter. Integration with the power conditioner, thermal control subsystem, and other interfaces would be resolved prior to a final transmitter design.

SECTION 5

DETAILED TECHNICAL RESULTS

Detailed considerations of the circumstances yielding the results of Section 3 are included here to show the technical development of the program. The general format follows that of Section 3.

5.1 TRANSMITTER SYSTEM DESIGN - TASK 1

This task evaluated the inputs, constraints, specifications, and interfaces among the other tasks as required to develop an optimum transmitter system. The results are distributed in the other Sections of this report to better provide an inclusive discussion of each part of the total system in a single place.

5.1.1 Overall Requirements

The overall requirements for the breadboard transmitter are in Appendix A. Since these are specifications and generally do not involve analysis from a system viewpoint, the details will be discernable in each of the task discussions. The individual task requirements are expanded where appropriate to include factors of particular interest to the specific transmitter system developed in this program.

5.1.2 Transmitter Design

The electrical and mechanical design of the transmitter can be found in the various sections of this report relating to the different subsystems. Sections 3.2.2 and 5.2.2 include the major mechanical design factors since the Doherty amplifier covered therein includes the mechanical assembly of most of the components of the system. System electrical design is also in those, and other pertinent, subsections.

5.2 VISUAL CHANNEL AMPLIFIERS - TASK 2

5.2.1 Visual Chain Driver Amplifier

5.2.1-1 Specifications

This task was to design, fabricate, and test a suitable driver amplifier for incorporation into the transmitter breadboard assembly. The driver amplifier is required to raise the power output level of the test exciter (5 watts) to a nominal 100 watt level required at the input to the Doherty visual power amplifier. The driver amplifier also must have essentially linear gain characteristics over the TV signal dynamic range and adequate bandwidth to avoid excessive distortion of the television signal. Dynamic bias control (remodulation) was not required for linearity correction with the ML8536 triode.

Specifications were included in Appendix A-2.1; additional design factors are:

Electrical

Load Characteristics	Load varies as a function of drive level
Plate Voltage Supply	1500 volts (maximum)
Circuit Configuration	DC grounded grid
Test Points	Monitoring points for all significant currents and voltages including rf input and output cavity voltage will be provided.

Thermal

ΔT between adjacent tube seals	100^{o}C maximum

Mechanical

Cavity Construction	Avoid excessive weight Avoid excessive thermal detuning effects (or provide for later incorporation of this feature.)
Auxiliary Circuit Construction	Package bias or other similar circuitry in a neat fashion. From the standpoint of personnel protection, this circuitry may be mounted as a subassembly in the test power rack, but must be designed so that it can readily be removed upon completion of the contract.

RF Connectors	Type N Coaxial
Power Connectors	No exposed voltages greater than 24 volts rms above reference ground. Connector design should permit ease in removing cavity from test setup and dismantling of the cavity.

Personnel Safety

High Voltage	All terminals more than 24 vrms above ground will be adequately insulated or shielded to prevent accidental contact by personnel.
RF Radiation	The level of all electromagnetic fields will be maintained below 10 milliwatts per square centimeter at all points accessible to personnel.
Hot Spot Temperature	All points on the outside surfaces of the circuitry which operate at temperatures above 100°C will be adequately shielded to prohibit personnel contact insofar as practicable.

5.2.1-2 Tube Selection

Several tubes were evaluated for the driver stage of the transmitter; Table 5-1 below shows the three most promising.

TABLE 5-1. PREFERRED TUBES FOR VIDEO DRIVER STAGE

Number	Type	Mfgr.	P_o KW	Class B N-%	BW MHz	Gain dB	Comments
ML-8534	Tri.	Mach.	0.32	51	31	15	Conduction Cooled, $\lambda/4$ circuits
ML-8536	Tri.	Mach.	0.18	49	29	15	Externally same as ML-8534
8226	Tet.	RCA	0.11	34	8	--	Plate $\lambda/4$, Cathode $3\lambda/4$

The 100 Watt or greater level for the driver stage provides a conservative operating power margin for driving the Doherty's two Y2042 or Y1498 tubes with normal circuit losses. The Machlett ML-8534 is a preferred candidate. A similar tube, the ML-8536 would be suggested for lower driving levels. Both tubes are conduction cooled and have good efficiency and bandwidth compared to competitive tubes. General Electric's

developmental Y1774 has a higher operating temperature capability which is advantageous for thermal radiation from the spacecraft, but little data on the tube is available and lack of assurance on its availability and consistency led to its rejection. Limited data available on the ML-8534 indicates a life expectancy of greater than 5000 hours in applications roughly comparable to that in this transmitter. Thus, several switchable driver stages may be employed with this driver tube for a long life mission. The ML-8534 was found to be power-level-sensitive to plate circuit loading; this situation is described in Section 5.2.1.5 on driver testing. The lower powered ML-8536 was substituted and provided adequate output power without encountering the problem of the other tube.

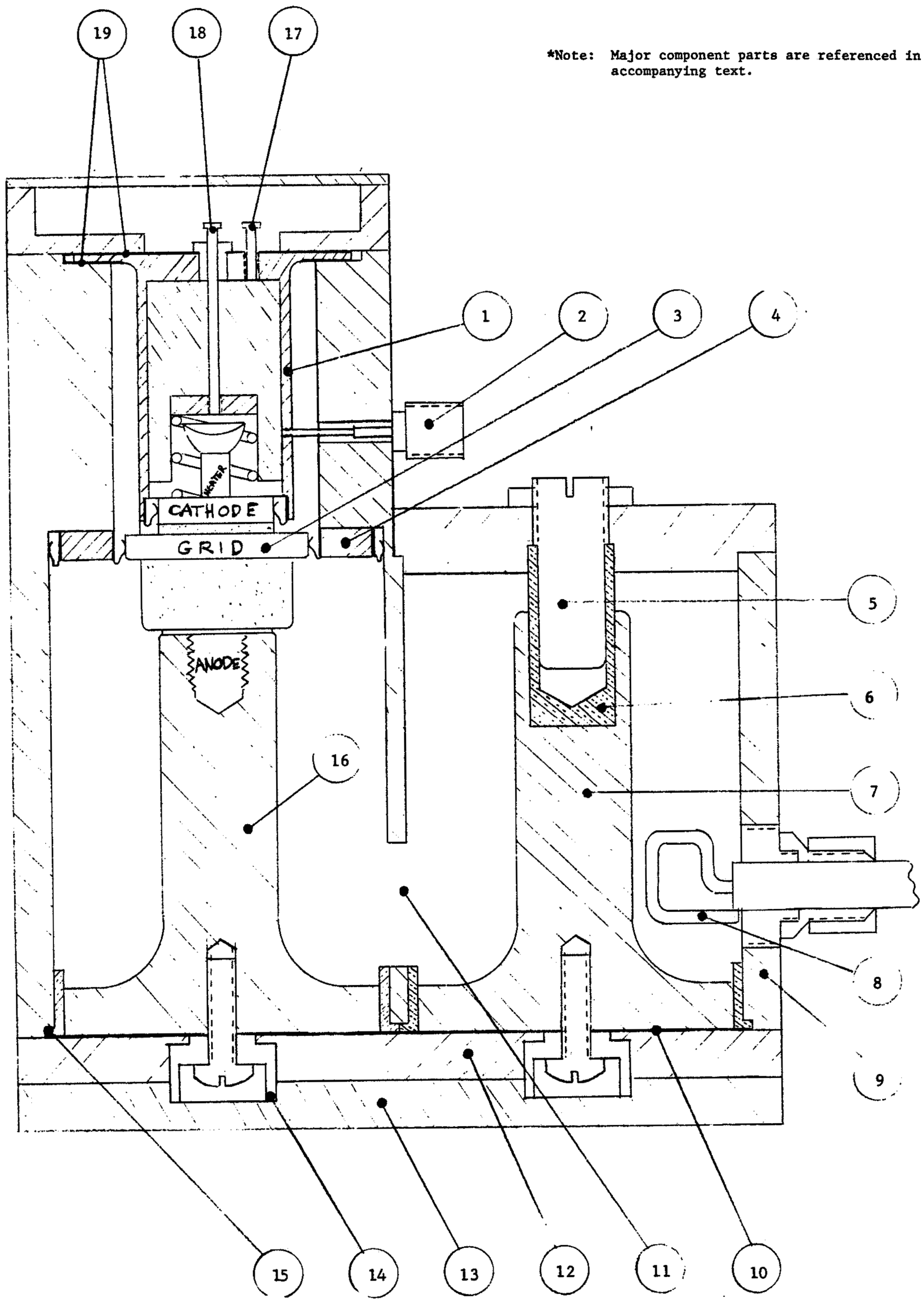
5.2.1-3 Design

The driver uses a grounded grid circuit as shown in Figure 5-1 . The grid is at dc ground as well as rf ground. This assures minimum rf feed-through and also aids in multipactor suppression. The cathode circuit is resonated by a low impedance short-circuited transmission line of less than one quarter wavelength. In Figure 5-1, the triode is item 3; the cathode line formed by the case and item 1 is rf shorted by the mica bypass capacitors, items 19. The input power is coupled to the line by a tap at a nominal 50 ohm point (item 2).

The heater voltage is brought in through the center of the cathode line to avoid the RF field. The cathode is common to one side of the heater which is contacted at the cold rf end at item 17. The plate circuit is also a quarter wave short circuited line. The line impedance chosen is a compromise between a small diameter center conductor to minimize the loaded Q and large diameter center conductor to minimize the temperature drop between the anode and case.

A second resonator is coupled to the plate line by an iris (item 11), thus forming a double tuned circuit. A double tuned circuit was necessary to obtain the required

*Note: Major component parts are referenced in the accompanying text.



Cross-section of Visual Driver Amplifier Cavity

FIGURE 5-1.

bandwidth for negligible effects on the transmitter passband.

The computed performance of the ML-8534 is as follows:

Plate Supply Voltage	1.2 kV
Grid Bias	-15 volts DC
Drive Power	6.37 watts
Input Impedance	75 ohms
Output Power	137 watts
Load Impedance	4.05×10^3 ohms
Plate Current	168 ma.
Grid Current	37.5 ma
Gain	13.3 dB
Efficiency	64%

Tube and Cavity Computed Characteristics

Effective Plate Tank Capacitance 3.35 pf
- this is Cgp (2.25 pf) increased 50% due to energy stored in transmission line

Loaded Q = 72.5

1 dB bandwidth single tuned	5.9 MHz
1 dB bandwidth double tuned	18.6 MHz

5.2.1-4 Multipactor Suppression Factors

Both plate lines in Figure 5-1 are operated at 1.2 kV above the case; this will normally prevent multipacting. The gap between the case and both plate lines will experience a peak RF voltage stress of 1.05 kV over 0.55 inches. This would cause multipactor breakdown in a vacuum environment if the gap had no dc bias. The gap between the cathode line and the case will have a maximum RF voltage stress of only 30 volts peak which is below the multipactor threshold.

5.2.1-5 Test Results

Initial testing was performed with the ML-8534 triode. In cold tests of the amplifier, the cathode cavity resonated at 875 MHz while the plate resonated at 819 MHz. A tuning capacitor between the case and cathode line near the cathode flange of the tube was adjusted, the plate line Z_0 was raised slightly, and the slug tuner was put into place to allow tuning of these circuits to the intended operating frequency (visual carrier = 825.25 MHz). There was no detectable rf leakage measured on the radiation survey meter at output power levels near 100 Watts. These initial tests also indicated an input VSWR of 1.8 and excessive output loop reactance for optimum plate loading, both of which were subsequently modified.

A serious problem was noted in these initial tests, however, in the form of the output loading effect as mentioned in Section 5.2.1-2. It is most clearly presented in swept frequency testing as shown in Figure 5-2. As plate circuit resonance is approached, a sharp reduction in output occurs and holds at this level (approximately)

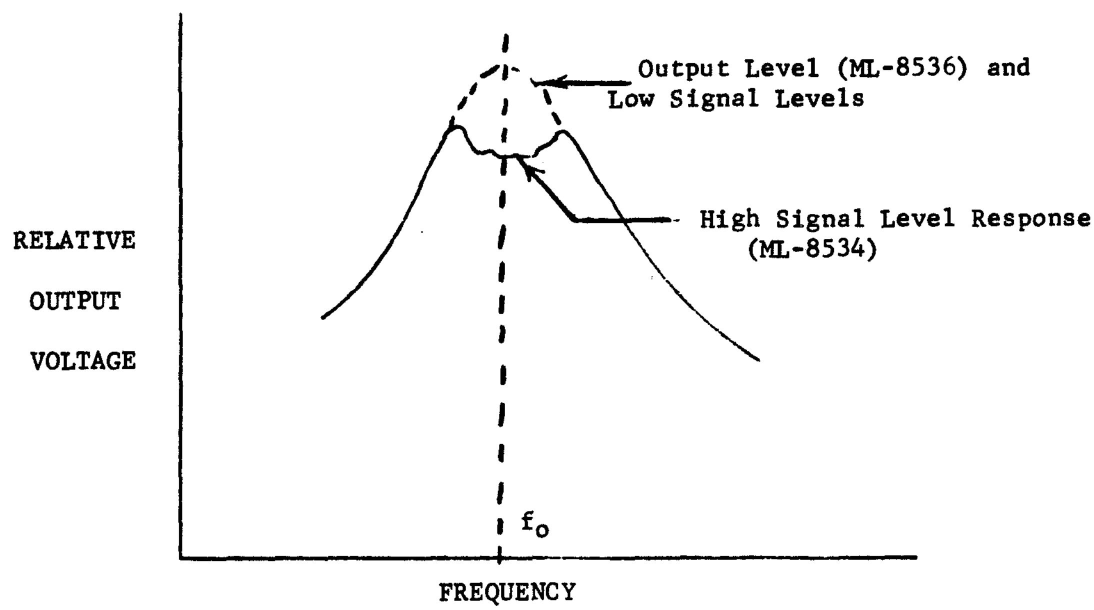


FIGURE 5-2. DRIVER PASSBAND

as the input drive frequency moves through the plate circuit resonance frequency f_o. This only occurred with sufficient drive to produce rated power output. A similar test at lower drive levels produced a normal resonance curve as indicated by the dashed line in the sketch (Class B bias conditions were used for the test). Since no evidence of oscillation at spurious frequencies was noted, it was speculated that some form of multipactor breakdown, occurring within the ML-8534 triode, may have caused the observed phenomena, which was exhibited in both of the available ML-8534 tubes. It should be recognized that multipactor breakdown outside the vacuum envelope was not possible in this test since tests were performed in the atmospheric pressure of a laboratory environment. Multipactor breakdown cannot occur except at high vacuum levels where collisions of electrons in the multipactor discharge with residual gas molecules are relatively infrequent.

The decision was then made to install a mechanically identical (external dimensions) ML-8536 triode as a means of identifying the nature of the problem. No evidence of this "breakdown" phenomenon was noted with the ML-8536. Since adequate power output and other parameters were obtained, the decision was made to utilize this tube in the driver for breadboard TV transmitter testing. Test performance data with the ML-8536 tube in service are included in the results Section 3.2.1-2.

The excellent amplitude linearity characteristics of the driver are illustrated in the TV waveform monitor stair-step display shown in Figure 3-4 and the corresponding differential gain display of Figure 3-5. The usual practise of video gain/phase precorrection was included within the TV exciter for these tests.

5.2.2 Doherty Amplifier

A high efficiency visual amplifier is required in the transmitter breadboard to provide a 5.0 kw output at the transmitter output terminal. The amplifier must have essentially linear gain characteristics over the TV signal dynamic range and adequate bandwidth to avoid excessive distortion of the television signal. Some degree of dynamic bias control (remodulation) is desirable for linearity correction and grid current limiting. A suitable amplifier was fabricated and incorporated into the transmitter breadboard assembly, where it was tested for operation in a high efficiency TV linear amplifier chain. A Doherty-type of linear amplifier was used since it would result in considerably higher efficiency than with the conventional Class B linear amplifier or currently available microwave type tubes operated as linear amplifiers.

5.2.2-1 Requirements

The major specifications for a Doherty High Power Amplifier are included in Appendix A-2.2. Following are some of the basic and supplementary specifications as derived in the Task 1 Systems Study:

Electrical

Frequency	825.25 MHz
Power Output	5.0 kw peak sync
Load Characteristics	1.2 (maximum) VSWR load
Efficiency Objective	>60% at 25% of rated peak output; 65% at peak sync
Plate Voltage	2500 volts (nominal)
Circuit Configuration	Based on Doherty Amplifier Circuit
Test Points	Monitoring points for all significant currents and voltages including rf input and output cavity voltages

Thermal	
Cooling Method	
Anode	Water or conduction (to heat pipe interface) depending on tube type and test fixture availability
Other tube and cavity elements	Conduction or radiation (no forced air)
Mechanical	
Cavity Construction	Breadboard design should be adaptable to space-type hardware with minimal changes. Design features should include: Ruggedness Avoidance of excessive mechanical stresses on the tube and other amplifier components due to external forces encountered during normal use, including thermal expansion. Avoid excessive weight. Cavity should be readily dismantled for tube replacement, developmental changes, etc. Avoid excessive thermal detuning effects or provide for a compensation feature.
Auxiliary Circuit Construction	Package bias or other similar circuitry to a neat fashion. From the standpoint of personnel protection, this circuitry may be mounted as a subassembly in the test power supply rack; however, it must be designed so that it can be readily removed for delivery to the customer upon completion of the contract.
RF Connectors	
Input	Type N Coaxial
Output	Mate with half-height WR 975
Power Connectors	No exposed voltages greater than 24 volts rms above reference ground. Connector design should permit ease in removing cavity from test setup and dismantling of the cavity.

Compatibility with Breadboard Circuit	Designs should be periodically reviewed with the project engineer to assure compatibility with all electrical and mechanical interfaces in the breadboard circuit.
Personnel Safety	
High Voltage	All terminals more than 24 volts rms above ground will be adequately insulated or shielded to prevent accidental contact.
RF Radiation	The level of all electromagnetic fields will be maintained below 10 milliwatts per square centimeter at all points accessible to personnel.
Hot Spot Temperature	All points on the outside surfaces of the circuitry which operate at temperatures above 100°C will be adequately shielded to prohibit personnel contact insofar as practicable.

5.2.2-2 Amplifier Design

A. Approach

The general aspects of the Doherty amplifier were outlined in the Results, Section 3.2.2; the circuit shown in that Section (Figure 3-6) included the input and dynamic bias circuits. This discussion will provide the details of the factors involved in its design.

The mode of operation is described in detail elsewhere[2] but will be reviewed briefly, using the simplified circuit of Figure 5-3. At low input signal levels, up to one-half peak drive voltage (or one-fourth peak drive power), only the first, or carrier, stage operates, transmitting its power past the "passive" cavity of the second stage to the output load. The $\lambda/4$ line provides an impedance inverting function such that the Carrier Stage loading is optimum for high efficiency at the half-voltage point. As the input signal increases, the second, or Peak, stage adds power to the load. At the same time, the effective impedance of the second triode reflects through the $\lambda/4$ line to the first stage, effectively lowering the load impedance seen by its plate, and permitting it to supply still

more output power. The voltage and power relations over a range from one-fourth to full output power is such as to provide a nearly constant high efficiency, 60% to 65%, over that range. This is the feature which results in high AM-TV efficiency compared to any other amplifier.

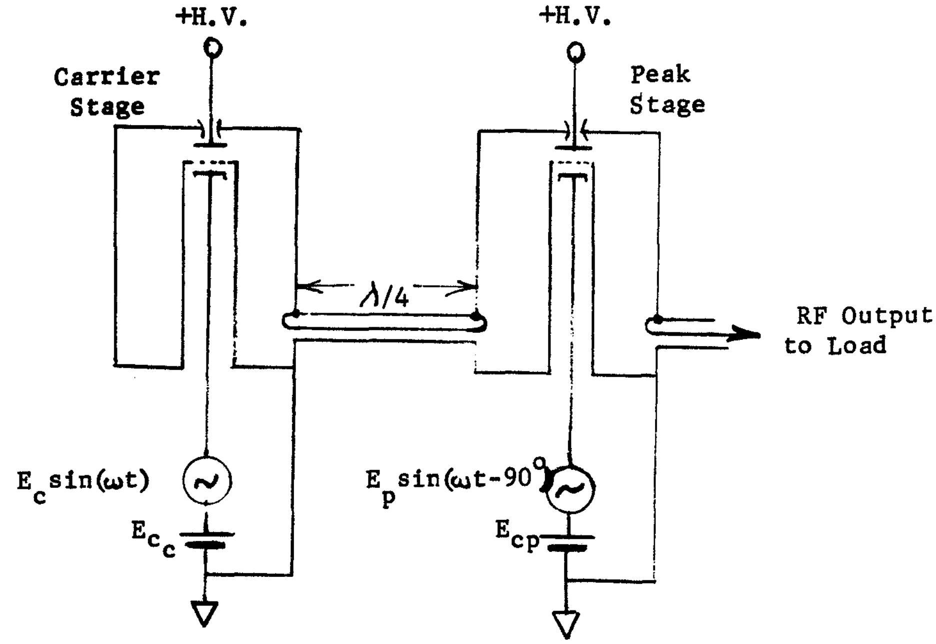


Figure 5-3. Basic UHF Cavity-type Doherty Amplifier Circuit

Two principal differences from the basic HF Doherty amplifier are included in this circuit. One is the low impedance couplings between cavities and between the Peak cavity and Antenna Load. The usual high frequency circuit is operated with a high impedance (quarter-wavelength transmission line equivalent) network between amplifier plates. Tests on a 30 MHz circuit, modeling the UHF cavity circuitry, demonstrated the feasibility of using low impedance couplings.

The second difference is the use of grounded-grid circuitry, which is the only feasible approach at UHF frequencies. In this arrangement, amplifier input impedance varies over a wide range, which must be considered in the design of the overall amplifier and driver circuit. In the grounded cathode circuit normally used in the HF frequency range, the problem of input impedance changes is solved by swamping and/or the use of a low impedance driver stage. Swamping is not acceptable for the grounded-grid arrangement since gain is lower than at HF and losses would be prohibitive. Also, the use of a low-source-impedance driver appears to be impractical since it is necessary to use interconnecting transmission lines between Driver and Doherty at UHF frequencies. Tests on the 30 MHz simulator* and other considerations indicate that appropriate design of the input power divider/phase shift circuit plus dynamic control of bias on the amplifier tube grids is necessary if reasonable distortion levels and high efficiency are to be achieved in the UHF circuit.

A simple block diagram of the visual signal chain with power levels and gains is shown in Figure 5-4. Bandwidth of the Y1498 is estimated to be about 55% of the lumped constant value of 19.2 MHz. This gives a response of about -0.5 db at the edges of the video passband. The Driver Bandwidth (from Section 5.2.2) is about 70% of the lumped constant value, or 21.7 MHz, which gives its response as about -0.3 db at the edges of the video band. Since these are the major bandwidth limiting factors, an adequate band margin exists. Some variation in bandwidth in the Doherty amplifier occurs over the dynamic range of the TV signal since both stages see varying impedance (though somewhat compensating) as drive level changes, but this effect is not significant; a cursory examination of delay relations indicates the differential phase due to the approximately 0.3 db reduction in bandedge response is only about 1^{o} at the color subcarrier frequency.

*Reference 1, Appendix A

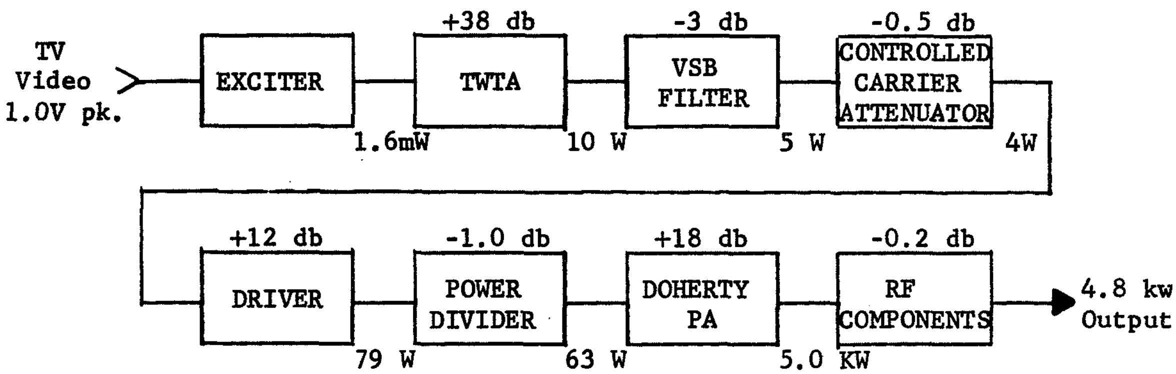


Figure 5-4. Visual Channel Signal Chain

B. Tube Selection

A review of amplifier tubes for the power amplifiers required in the transmitter was made. Comparisons were made on the basis of published or calculated efficiencies, calculated bandwidth, gain, circuit implementation factors, and cooling system requirements. Where necessary, estimates were obtained from calculations based on constant-current data curves for that tube, using the Fourier analysis method published by Eimac division of Varian[4].

Single tuned 3 db bandwidth was estimated from the lumped constant (or "low frequency") method which is based on the load impedance R_L and output capacitance C_o:

$$\Delta f = \frac{1}{2\pi C_o R_L}$$

Here, R_L is set primarily by required power output and maximum tube ratings, while C_o is taken from the tube data sheet. (In precise calculations, the actual tube cross-section must be analyzed to correct C_o for tube parasitic element values.)

Bandwidth requirement is based on allowing 0.5 db variation in response over a 5 MHz amplifier passband, which is equivalent to a 3 db variation over about 14.3 MHz.

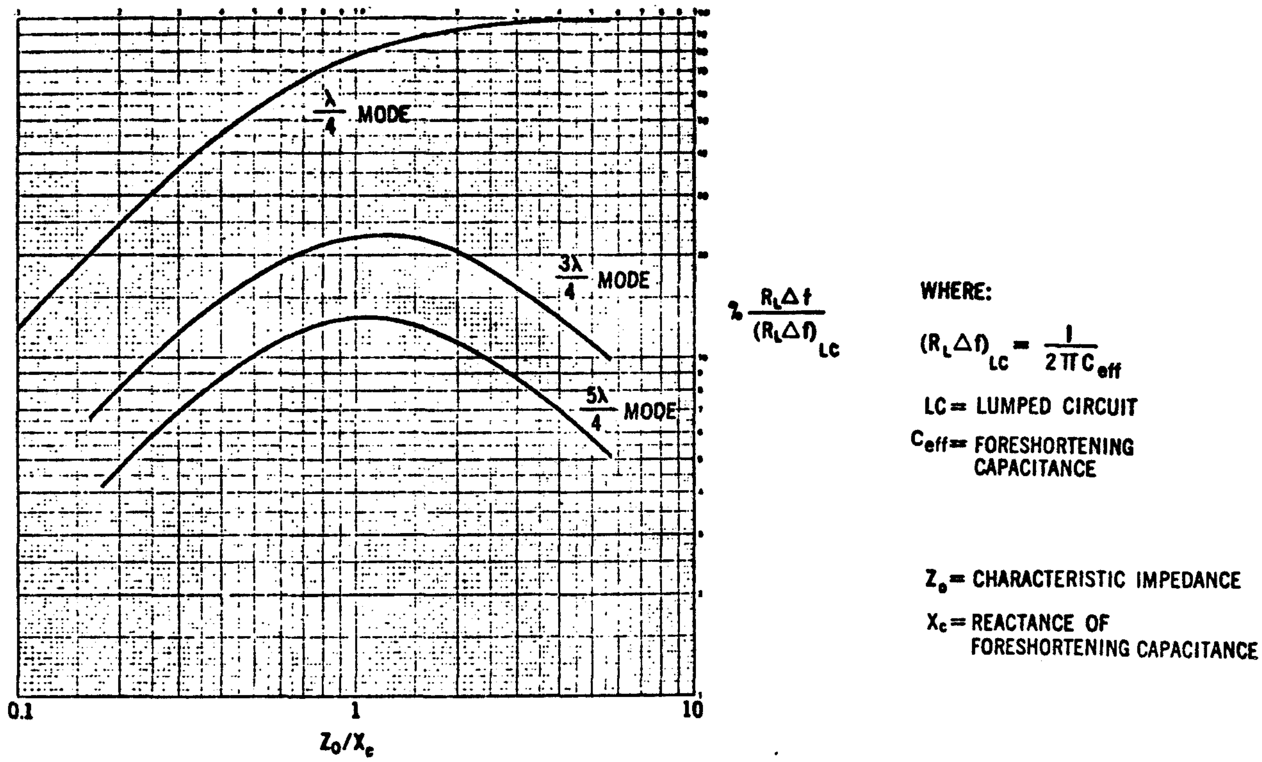
Generally, it is the output circuit of a grounded grid amplifier rather than the input circuit that is the constraining item in determining amplifier bandwidth. Two additional factors were considered in the amplifier design: the bandwidth of a tank circuit utilizing transmission line elements is always less than the equivalent lumped-element circuit bandwidth, and additional 0.5 db bandwidth is possible with a double tuned circuit.

Bandwidth reduction is shown in Figure 5-5(a)[4] for cavities operating in the $\lambda/4$, $3\lambda/4$ and $5\lambda/4$ modes. Typically, a quarter-wave cavity will have about 70% of a lumped constant circuit bandwidth. Clearly, the $3\lambda/4$ cavity, which must be used with physically large tubes, should be avoided since it reduces bandwidth to less than 25% of the lumped constant value. Double tuning can increase bandwidth substantially, more than doubling the 0.5 db bandwidth as indicated by the data in Figure 5-5(b). Double tuning is included in the driver design in Section 5.2.1.

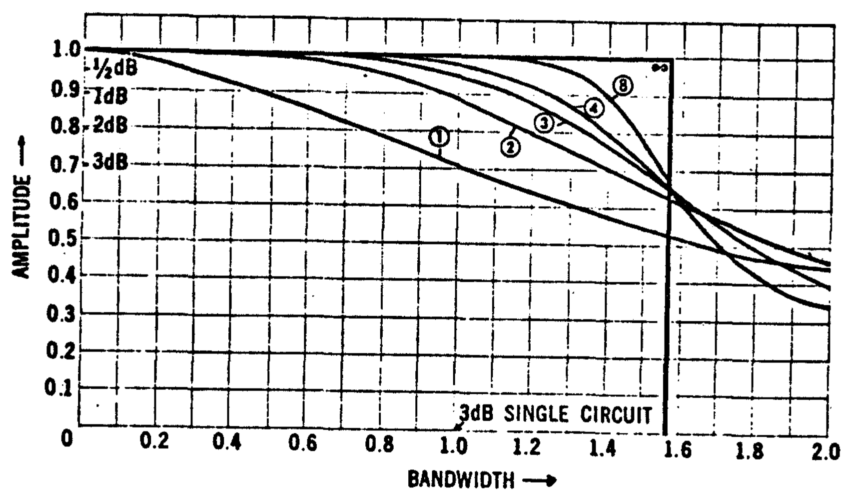
Over twenty tube types were included in the tube review. Of these, the best are included in Table 5-2 below.

Table 5-2. Preferred Tubes for High Power Stages

Number	Type	MFGR.	P_o kw	Class B η - %	BW MHz	Gain db	Comments
Visual PA							
Y2042, Y1498 (L-64S)	Tri.	GE	2.5 / 1.0	61	31	19 / 17	$\lambda/4$ Input and Output
GL6942	Tet.	GE	1.0	55	12	10	$\lambda/4$ Output Circuit
7213	Tet.	RCA	1.3	54	13	13	$3\lambda/4$ Grid and Plate Lines



(a) Bandwidth Reduction Factors for Transmission Line Resonators



(b) Bandwidth Relations for Multiple Tuned Resonant Circuits

Figure 5-5. Bandwidth Relations for RF Amplifiers

The preferred tube for the Doherty is clearly the Y-2042, a high performance version of the Y1498; these two types are both production versions of the L-64S tube but the Y1498 uses an oxide coated cathode which gives it a shorted lifetime than the five-year (predicted) Y-2042. Unfortunately, the Y-2042 was not available as an operational tube during the performance period of the contract and the Y-1498 was used. However, this series still provides the best power output, efficiency, gain, and bandwidth. The Y-2042 (or Y-1498) is adaptable to conduction cooling with heat pipes at high operating temperatures, a feature not offered by the GL6942 and 7213. The same tube is also a good selection for the aural amplifier, discussed in Section 5.3.

C. Doherty Amplifier Characteristics

Operating conditions of the two amplifier stages of Figure 5-3 must be set so that essentially linear amplifier transfer characteristics are obtained and that the high efficiency characteristics of the Doherty circuit are realized. Complete characterization and analysis of circuit performance over the linear range is very complex due to the non-linear characteristics of Class B and C amplifiers and the interrelations between the carrier and peak stages of the amplifier. Performance characteristics were therefore considered at a few discrete points, and the circuit was adjusted experimentally to give acceptable results. Characteristics of the two stages at various points of input operating voltages are summarized in Table 5-3:

Table 5-3. Operating Characteristics of Doherty Stages

Condition	Carrier Tube	Peak Tube
1. Drive voltage below carrier level (less than one-half peak)		
a. Bias Point	Idling current point set for optimum compromise between high efficiency and best linear operation at low levels (typical Class B condition).	Bias greater than cut-off value. Plate voltage swing is 0.707 of carrier tube swing due to carrier stage feeding through peak cavity to load.

Table 5-3 (Cont'd)

Condition	Carrier Tube	Peak Tube
b. Plate load impedance	Twice that at full peak power.	Not applicable.
c. Input impedance	Higher than for peak levels; varies at low drive levels.	Very high.
2. Drive voltage at carrier level (one-half of peak)		
a. Bias point	Combination of bias and rf drive voltage to give full allowable plate swing (peak voltage swing 80 to 90% of plate supply voltage). Saturation conditions occur.	Combination of bias point and rf voltage causes small conduction.
b. Plate load impedance	About twice that at full peak power. Impedance level begins to decrease as peak tube starts to deliver power to the load.	High, with voltage induced by current in load from carrier stage and a small current in peak stage.
c. Input impedance	Higher than for peak level operation. Begins to decrease as tube is driven above carrier operating point due to greater loading of plate circuit as Peak Stage delivers power to load.	Very high. Decreases as peak stage begins to deliver current to load.
3. Drive voltage at peak level		
a. Bias point	Stage is normally heavily saturated; dynamic bias preferred here to alleviate undesirable condition of overdriven grid.	Peak stage is delivering same power to load as carrier stage since instantaneous peak drive voltages are roughly the same. Stage is operating at saturation point.
b. Plate load impedance	Rated value for full peak power.	Rated value for full peak power.
c. Input impedance	Design value for peak power operation.	Design value for peak power operation.

The nominal operating characteristics of the Y2042 from experimental tubes will be as follows for a 5.0 kw sync peak output:

	Carrier Tube	Peak Tube
Plate Voltage	2.5 kV	2.5 kV
Plate current at sync peak	1.5 A	1.4 A
Grid Bias (no signal)	-12 V	-30 V
Grid current	0.5 A	0.5 A
Peak rf grid volts (above zero)	+6 V	+7 V
Plate dissipation	1.25 kW	1.0 kW
Grid dissipation	1.7 W	0.7 W

D. Anode Cavity Design

The Doherty cavities are rectangular, radial line types, rather than round; these ease tuning adjustments because of better physical compatibility with the mating waveguides used for the high-power transmission lines and for the interstage quarter-wavelength coupling line. The amplifier circuit is an rf and dc grounded grid configuration as was shown in Figure 3-6. The anode bypass is a multi-stage low rf leakage design. (This bypass also serves as a model for future dc grounded plate cavity designs where very low leakage between cathode (input) and anode (output) cavities is required.)

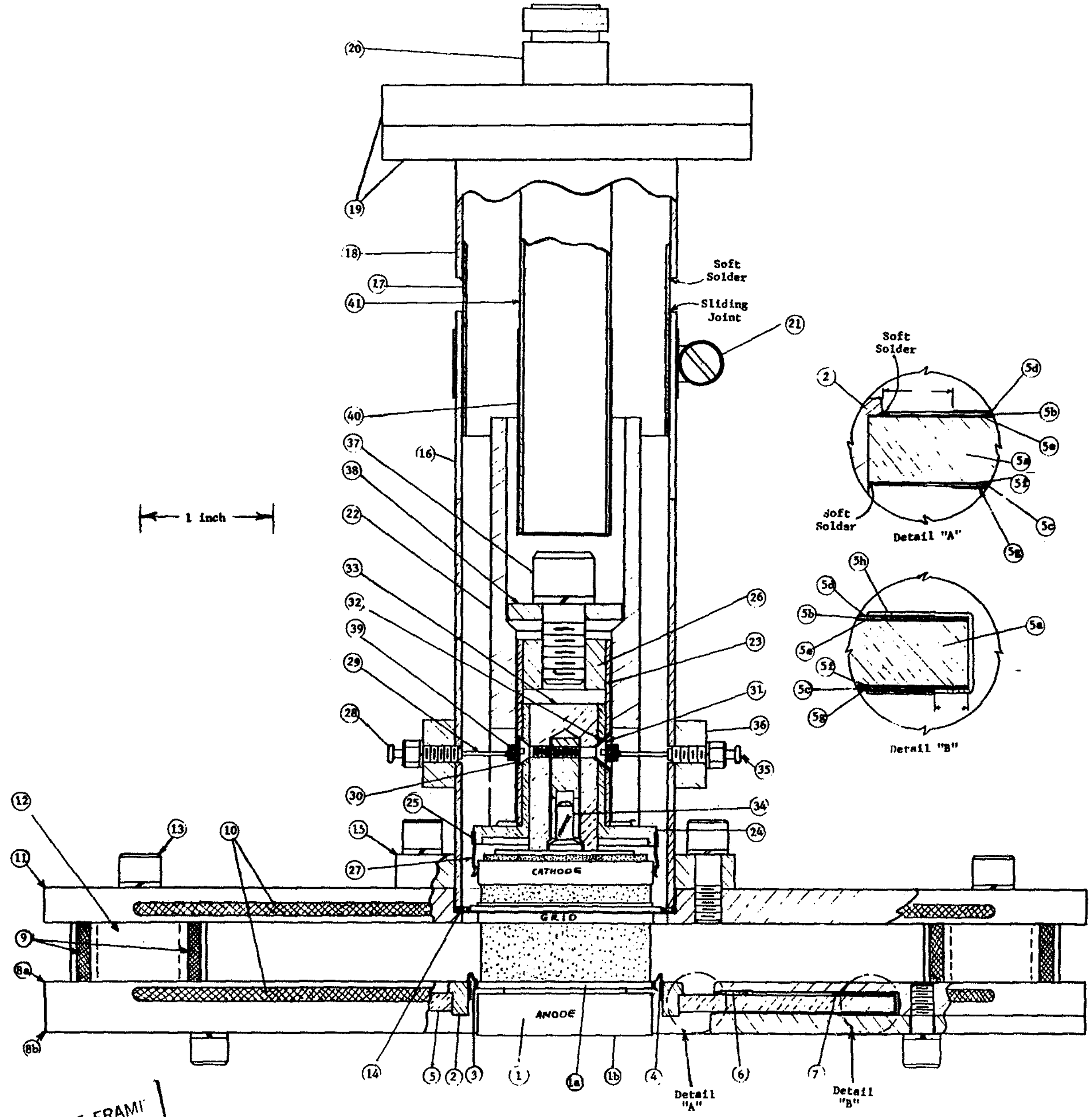
The input impedance of the cavity is a variable magnitude, variable phase load which requires special treatment to avoid excessive differential phase characteristics in color TV signals. The cavity as designed gives essentially a resistive input impedance over the range encountered as the drive signal varies from minimum to maximum.

Other features of the cavity include a resilient conductor having good electrical, thermal, and low outgassing properties for grid and bypass capacitor rf contacts, a water-cooled heat sink at the conduction cooling interface of the Y1498 anode, and monitoring points for input and output cavity rf phase.

An axial cross-section of the amplifier cavity is shown in Figure 5-6. The 5.30 inch square anode radial cavity is equivalent to the 2.79 inch radius round cavity used for the aural stage amplifier, Section 5.3. Two opposing sides of the square cavity are omitted to form the coupling irises to the mating waveguides for the Doherty peak amplifier. In the carrier amplifier, an identical cavity is used with one of the coupling irises replaced by a tunable short, as shown to the left of the cavity cross-section. The remaining two sides of the cavity are adjustable over nearly a one-inch range to permit resonant frequency tuning adjustment. Square Teknit* resilient gasket material is used to allow sliding adjustment of these side pieces under high power operation with moderate fastening screw tension applied. The same Teknit material is used in the mating surfaces of the cavity to the waveguide irises. The various components of the Doherty amplifier (cavities, iris plates, and waveguide) are clamped together by two long threaded rods; the resilient Teknit gasket provides uniform contact between the components with only moderate clamping pressure. Thus, the amplifier can be disassembled quickly for inspection and adjustment as required during the test process.

The plate bypass assembly is clamped between the two lower plates of the anode cavity in a 4.9 inch recess cut into these plates. This capacitor consists of three $\lambda/4$-wavelength radial-line sections arranged in the circuit of Figure 5-7.

*Trademark of Technical Wire Products Company, Cranford, New Jersey.



PARTS IDENTIFICATION

1. GE Y1498 planar triode
   a. RF contact
   b. Thermal interface, B+ connection
2. Anode contact ring
3. Anode contact fingers
4. Anode contact finger retainer ring
5. Anode bypass capacitor assembly (sandwich of following)
   a. Intermediate choke dielectric, custom Hi-K 707, double clad
   b. Input capacitor dielectric, ruby mica, silvered
   c. Output capacitor dielectric, ruby mica, silvered
   d thru g. Silver or copper conductors clad or deposited on items (a) and (b)
   h. Encapsulant on capacitor edge, GE Irrathene strip
6. Capacitor input contact, Chomerics 1224 conductive gasket
7. Capacitor output contact, Chomerics 1224 conductive gasket
8. Anode cavity lower plate
   a. Main plate
   b. Capacitor recess cover
9. Output coupling iris contacts for tuning bars, Teknit strip
10. Output coupling iris contacts for cavity top & bottom, Teknit strip
11. Anode cavity upper plate
12. Anode cavity tuning bars
13. Anode cavity fasteners, 10-32 Allen head screws & split lockwashers
14. Grid contact, Chomerics 1224 conductive gasket
15. Cathode cavity flange
16, 18. Cathode cavity outer conductor
17. Cathode cavity tuner, outer conductor
19. Cathode cavity input transition to Type-N coax connector
20. Input connector, Type N, 50 ohms
21. Cathode cavity tuner clamp
22. Cathode cavity matching transformer
23, 24, 25, 26. Cathode contact assembly
27. Cathode contact fingers
28. Cathode terminal/bypass capacitor
29. "High inductance" cathode contact lead
30. Heater insulator retaining screw
31. Heater contact retaining screw
32. Heater contact retaining screw insulator, 0.016" Teflon
33. Heater contact insulator, Teflon
34. Heater pin contact
35. Heater terminal/bypass capacitor
36. Bypass capacitor mounting block
37. Matching transformer retaining fastener, 5/16" Allen head
38. Matching transformer retaining washer
39. Heater and cathode resilient contacts, Teknit "fuzz button"
40. Input tuning capacitor protective insulation, GE Irrathene
41. Input tuning capacitor movable electrode

Figure 5-6. Axial Cross Section of Doherty Stage Cavity

FOLDOUT FRAME 1

FOLDOUT FRAME 2

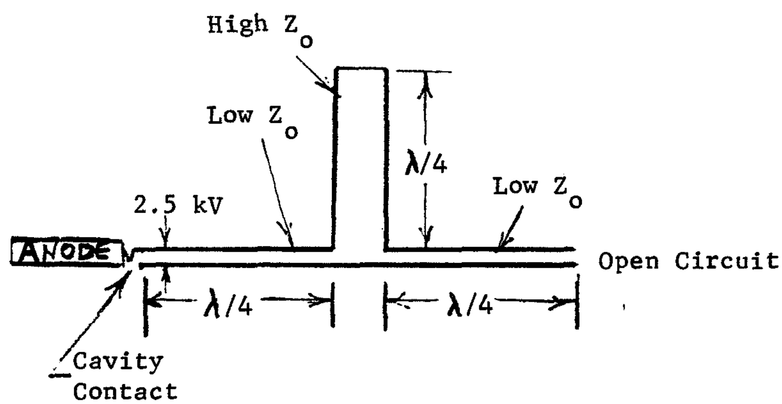


FIGURE 5-7 HIGH-ISOLATION BYPASS CAPACITOR ASSEMBLY

Silvered mica sheets of 2.5 mil thickness are used in the low Z_o sections; the high Zo section uses 1/8 inch thich Custom Materials 707K copper clad dielectric with a relative dielectric constant of 3.0. The sections are stacked as shown in Figure 5-6 to form a compact assembly. The plate contact ring is soldered to the inner edge while the outer edge is encapsulated with heat shrinkable Irrathene* to increase the high voltage creepage path length from the bypass to the cavity. Uniform electrical contact between parts of the assembly is obtained from rings of Chomerics type 1224 resilient silver/silicone rubber gasket material which press the conductive survaces of the $\lambda/4$ radial line sections together in narrow bands around their inner and outer perimeters. The inner ring also provides electrical and thermal contact between the capacitor and cavity.

* Trademark of General Electric Company

Input admittance of the cavity was determined using a three-probe impedance type of measurement over a range of operating conditions from plate current cut-off to simulated full power operation. This admittance consists of

$$Y_{in} = g_{m_{total}} + j\omega C_{in} \text{ (=0.25 + j 0.25 at peak rf drive)}$$

where $g_{m_{total}}$ is the sum of grid and plate transconductance, and C_{in} is the input capacitance of the tube which is mainly the grid-cathode capacitance. The resistive component varies over a wide range as amplifier drive is varied, particularly for the peak amplifier cavity. With the usual case of a matched impedance (resistive) signal source impedance, it is apparent that variations in transmitted phase will occur with drive level changes. A special input impedance matching section was designed to match the input impedance of the cavity to 50 ohms (Y = 0.02 + j 0) at carrier or peak level (as desired) while providing nominally Y = G + j 0 conditions at other drive levels, thus reducing differential phase variations due to this cause to an acceptable level.

The matching transformer design in indicated in Figure 5-8. The measured input impedance versus rf drive locus at the Y1498 cathode interface is indicated by curve #1. A 17.3 ohm matching line of 0.204 wavelength transforms this impedance to that of curve #2. This is simply adjusted to coincide with the resistance axis of the Smith Chart by the insertion of a capacitor with X_C = 57 ohms (curve #3). Thus, the variable impedance/phase input impedance is transformed to a variable resistance, fixed phase input impedance which is matched to 50 ohms at peak level in the example shown.

The implementation of the input circuit is also illustrated in Figure 5-6. Provisions are made for inserting various transformer line sections, and a

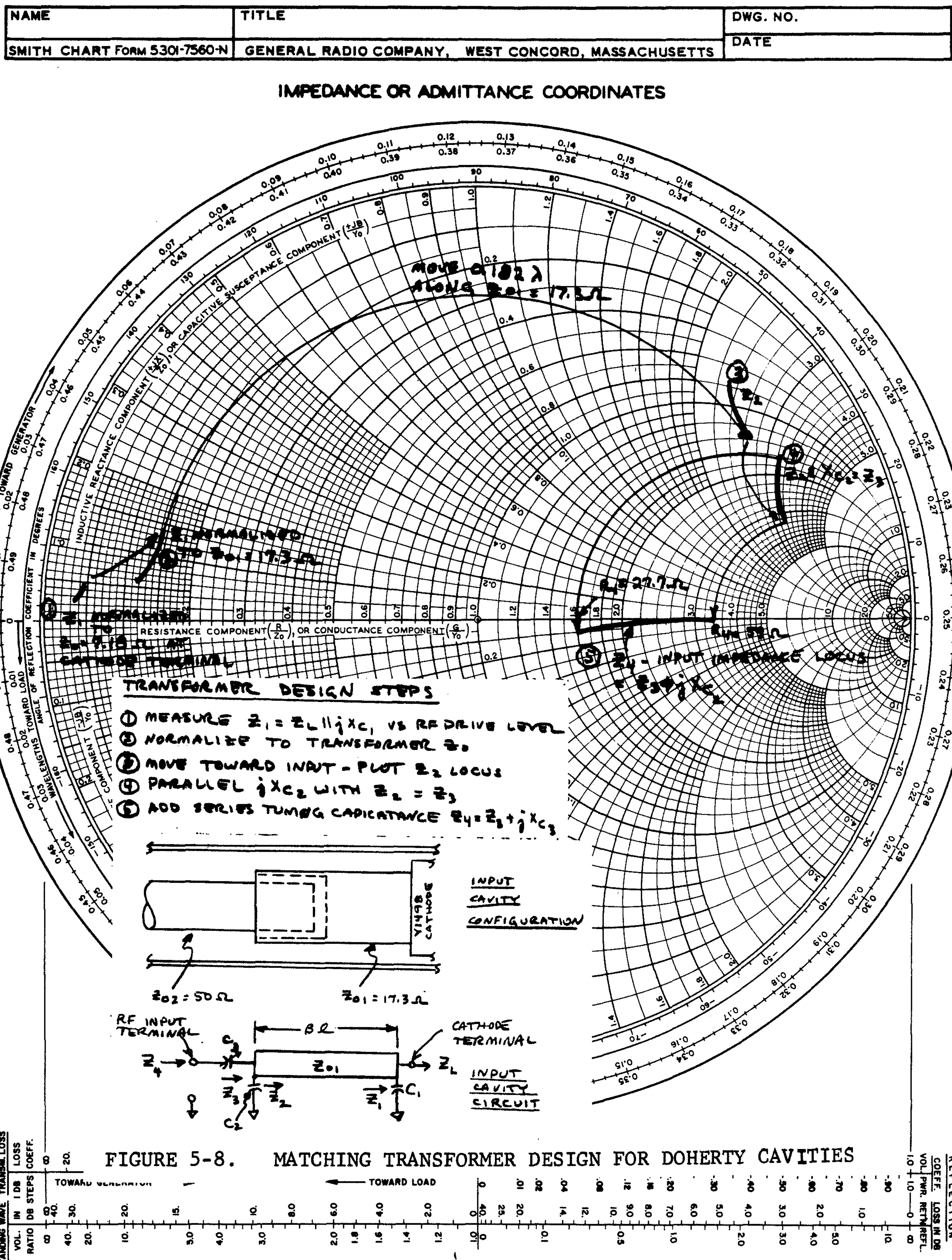


FIGURE 5-8. MATCHING TRANSFORMER DESIGN FOR DOHERTY CAVITIES

cylindrical series tuning capacitor is used to allow further adjustment flexibility. The input section outer conductor consists of two telescoping cylinders with a clamp which permits adjustment of series capacitance.

E. Doherty Cavity Performance

Preliminary measurements of the individual Doherty Amplifier cavity showed generally good performance with some problem areas noted. The results are summarized below:

• Input VSWR	Less than 1.5 over a 48 MHz bandwidth (corresponds to -0.2 db response)
• Input-Output Isolation	42 db
• Output Coupling	Inadequate ($\ll$ 1/3 that required for peak cavity loading)
• Input VSWR	Less than 1.5 over a 48 MHz bandwidth (corresponds to -0.2 db response)
• Output Bandwidth	13 MHz with full 5.3 inch iris width (share circuit on second iris port)
• RF Operation	Stable but power output and efficiency appear low
• Hum and Noise	A 0.15 volt drop in the heater contact assembly results in hum modification of rf output; will have to change to dc heater and/or reduce drop. Hum also noted due to poor driver HVPS output ripple.
• RF Leakage	No reading detectable on Sperry model B86B3 Survey meter.
• Harmonics & Spurious	Rough measurement indicates second harmonic -38db; third harmonic -50 db (including +6 db/octave loop coupling sensitivity factor).

The hum problem was reduced to a satisfactory level by changing the Y1498 heater supply and the driver power supply; in addition, some of the resistance in the lead common to the cathode and one side of the heater was reduced by rewiring.

An approach to increase coupling by tuning the iris with a shunt capacitive gap was investigated. A 0.70 pf capacitance made up of a 5/8 inch wide post was centered in the iris opening. However, in a mechanically suitable form for high power operation, a curious double-tuned iris coupling response was obtained. Under high power the higher loading (as evidenced by increased plate current) was not matched by power output; i.e., efficiency was low. Another approach of augmenting iris coupling by means of a coupling loop extending through the loop opening appeared more promising and was used to obtain some data at the Channel 73 carrier frequency of 821.25 MHz. With this loop, 400 watts CW output at 50% efficiency with a plate voltage of 1000 volts was obtained. (Previous efficiencies normally ran 25 to 40% and results were generally not repeatable.) Operation of the same cavity at a plate voltage of 1500 volts resulted in failure (puncture) of the plate bypass capacitor mica dielectric. This unit had previously been painted with a series of Tempilaq* temperature indicating paint. A very definite heating pattern was noted which indicated the presence of a peripheral resonance mode along the radius of the inner resilient contact ring with a maximum voltage at the point nearest the output coupling iris. The cavity was noted to be rather warm to the touch during operation.

Similar efficiency results were noted in the other cavity amplifier stage at f_o = 821.25 MHz. However, operation at 820 MHz with loop augmented coupling resulted in 63% Class B efficiency at 335 watts output with 985 volts on the anode. Unfortunately, operation at a lower frequency did not appear feasible without extensive rework of the Doherty circuit since many components were optimized and fixed tuned for UHF Channel 73.

*Trademark, Tempil Division, South Plainfield, New Jersey

The best prospect for proper operation appeared to be some rework of the Doherty amplifier cavities, but it was decided that some very pertinent information would be gained by first evaluating the performance of the Doherty amplifier. Data was taken on the two amplifiers in separate operation as a basis for establishing Doherty testing procedures as a basis for comparing conventional Class B and Doherty operation. Typical data representing the most significant operating modes is presented in Table 5-4. Television stairstep test signal photographs shown in Figures 3-12 and 3-14 are keyed to this data.

Future amplifier testing with approximately 10% change in capacitor inner diameter would be desirable to verify the amplifier design at the intended channel 73 operating frequency of 821.25 MHz. Doherty amplifier tests reported in the next section used the present amplifier with reduced output couplings to avoid possible damage to the bypass capacitor.

F. Mechanical Implementation

Implementation of the Doherty amplifier is illustrated in Figures 5-9 and 5-10. This assembly is mounted on the heat sink plate of the breadboard transmitter as shown in the photograph of Figure 3-2, repeated here. The plane of the assembly sketch, Figure 5-10, includes the axis of the interconnecting waveguide and the Y1498 amplifier input cavities as indicated in the assembly plan view of Figure 5-9. The assembly shown is clamped together with two sets of two threaded rods each, which are visible in Figure 3-2. This clamping approach permits quick disassembly for circuit adjustment or other work. In cases where less than full coupling iris width is required for anode cavity output loading, shim strips with the appropriate iris dimensions cut into them are simply slipped between the cavity and coupling iris plate,and the assembly clamped tightly against the shim for electrical contact. Anode and grid tuning are performed as described previously for the Y1498 amplifier cavity.

TABLE 5-4. PERFORMANCE OF SINGLE-STAGE Y1498 CLASS B AMPLIFIER

OUTPUT COOLING	MODE	E_{bb} Volts	E_{cc} Volts	Avg. Amperes I_B	Avg. Amperes I_g	Avg. Watts P_{in}	Avg. Watts P_{out}	η	f_o MHz	BW MHz	Y1498 Tube Serial #
Cavity 1											
5.3 in.x 0.4 in. iris + Capacitance	CW	960	5.2	.405	.240	5.6	165	42.5	821.25	13	Tube #135
	TV	1030	5.3	.275	.020	1.6	86	30.4	821.25	13	Tube #135
5.3 in x 0.28 in. iris + Capacitance	CW	1000	5.3	.230	.035	1.31	58.5	25.2	821.25	6.3	Tube #135
	TV	1000	5.3	.115	.005	.045	21.0	18.2	821.25	6.3	Tube #135
Cavity 2											
5.3 in. x 0.42 in. iris + Capacitance	CW	1000	5.3	.340	.090	3.3	143	42.0	821.25	10.6	Tube #136
	TV	1000	5.3	.210	.010	1.1	65	31.0	821.25	10.6	Tube #136
5.3 in. x 0.28 in. iris + Capacitance	CW	1000	4.3	.215	.015	.043	52	22.4	821.25	7.8	Tube #136
	TV	1000	4.3	.155	.000	0.11	19.5	12.6	821.25	7.8	Tube #136
5.3 in. iris + 0.4 in^2 loop	CW	985	5.8	.570	.240	---	355	63.3	820	---	Tube #136
5.3 in. iris + 0.5 in^2 loop	CW	1000	5.8	.680	---	---	375	55	803	10.5	Tube #136

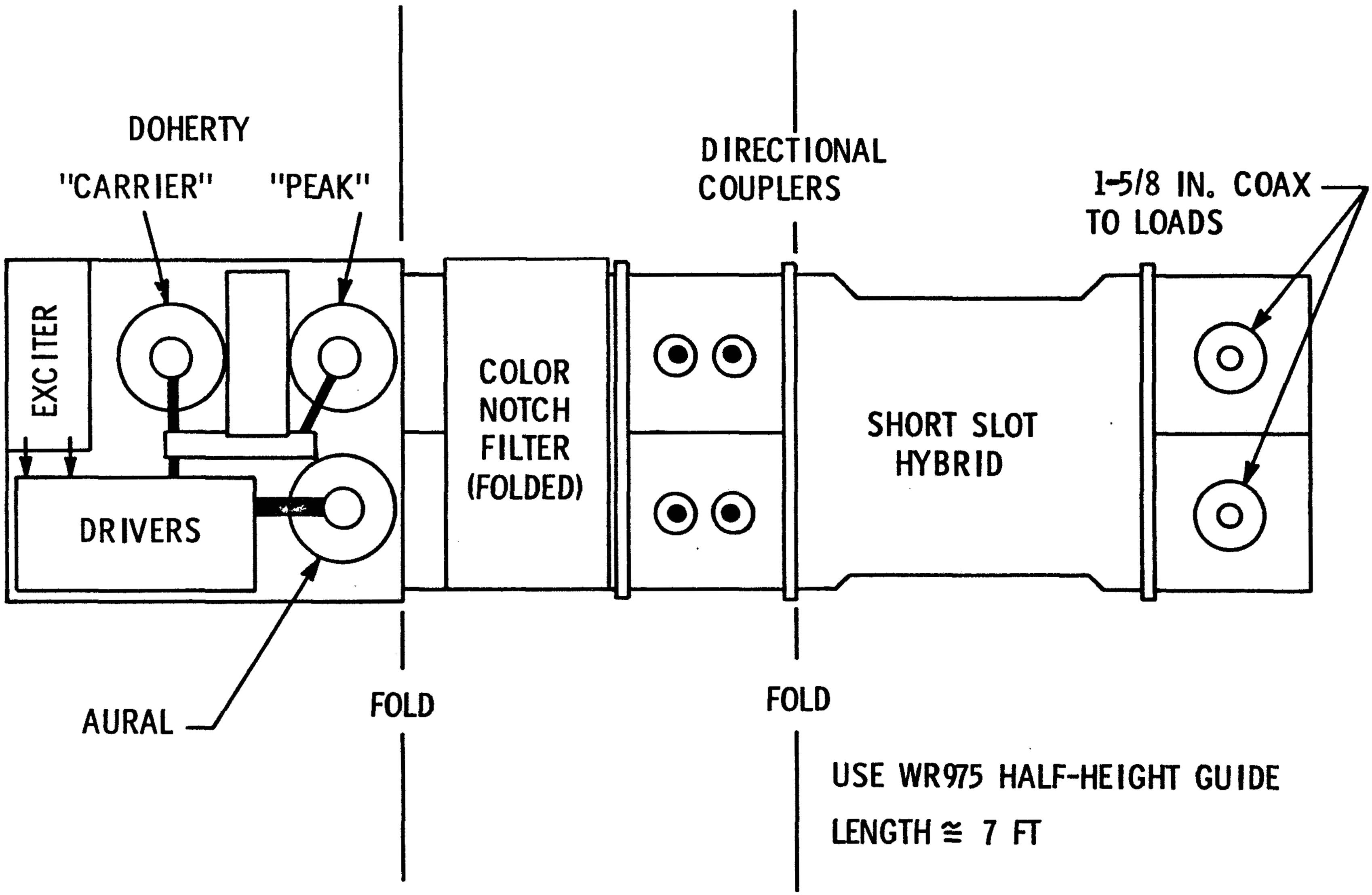


Figure 5-9. Breadboard Transmitter Mechanical Layout

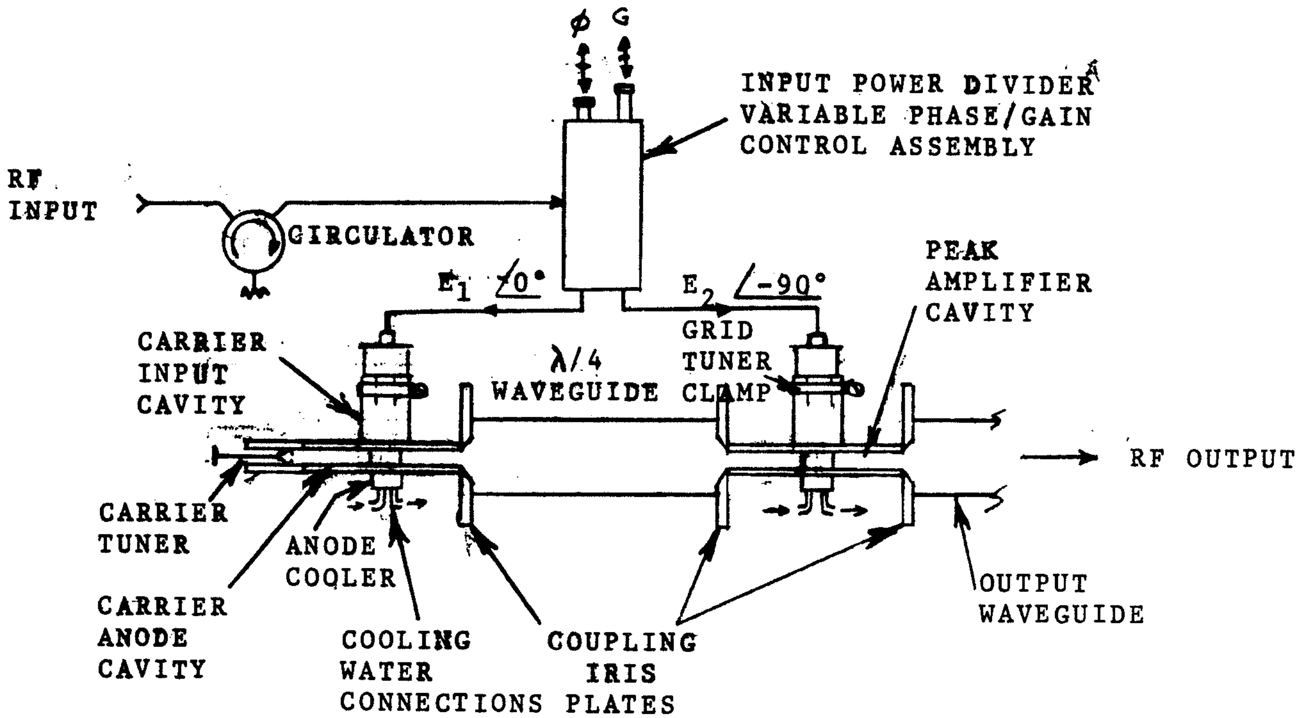


FIGURE 5-10. DOHERTY AMPLIFIER ARRANGEMENT

A coax assembly is provided which divides the rf input signal from the driver to the two Doherty stages, allowing for adjustment of relative peak input signal phase and carrier input signal amplitude. The circuit for this unit is shown in Figure 5-11. Hybrid #1 gives a 3 db power split and provides the -90° phase shift for the peak cavity. Hybrid #2 has shorts on the coupled parts which are ganged to move in the same plane, thus acting as a phase shifter for the peak stage. Hybrid #3 has ganged shorts which move in opposing directions, causing this part of the circuit to act as a variable attenuator or gain control for the carrier stage. Signal reflections from the input circuit

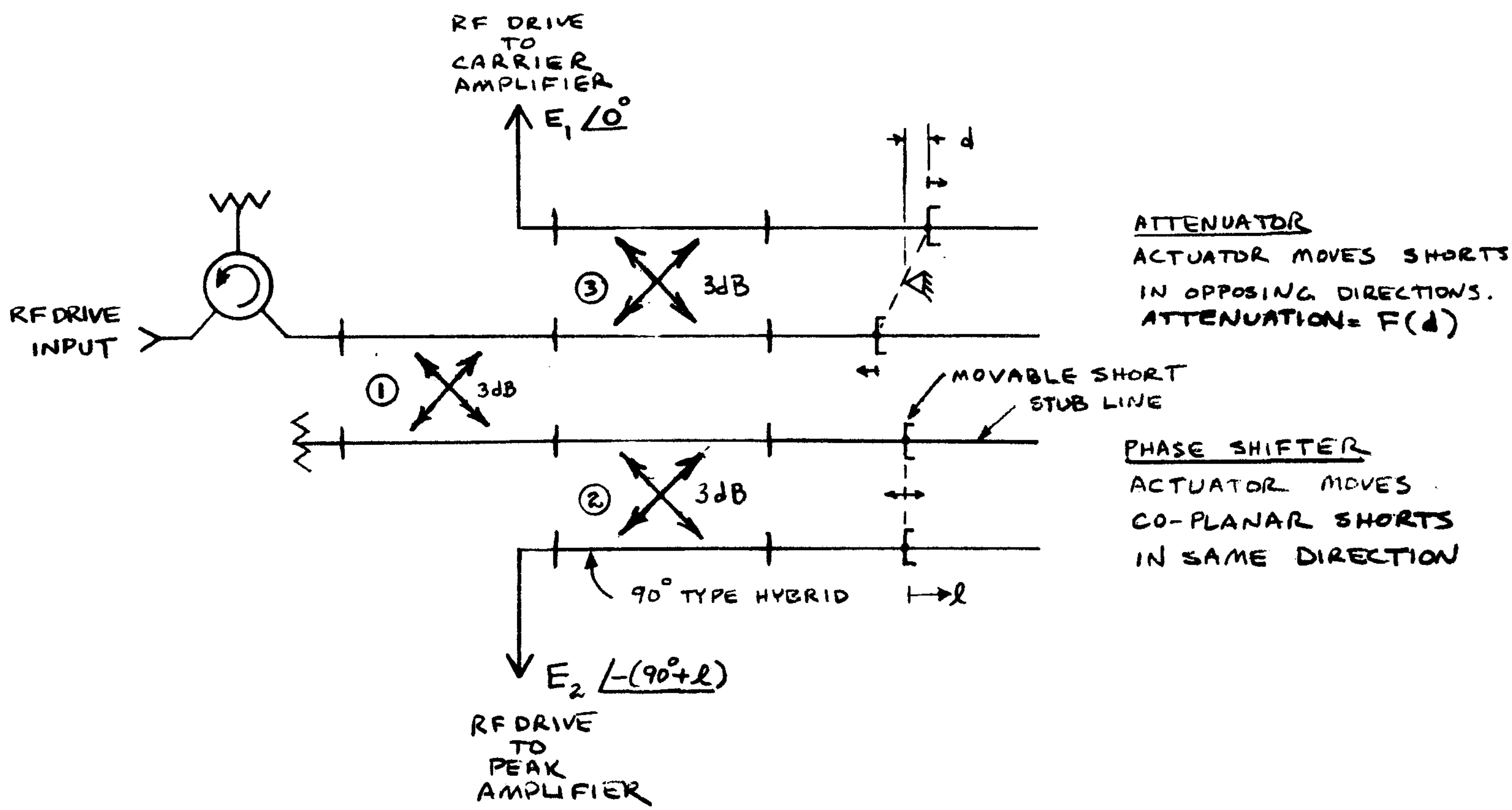


FIGURE 5-11. INPUT SIGNAL DIVIDER

are absorbed in the circulator and Hybrid #1 terminations. Line lengths and other input circuit elements are chosen so that proper phasing is approached with the Hybrid #2 shorts placed in the neutral or centered position. The input assembly is mounted over the $\lambda/4$ waveguide section as shown in Figure 3-2 .

G. Tuning

A tuning procedure was devised which allowed most adjustments to be pre-set on the individual cavities before coupling them together in the Doherty amplifier circuit. This procedure is outlined in the following steps:

1. Characterize individual amplifiers feeding a matched waveguide load. Note tuning and loading settings.

2. Set up Doherty amplifier circuit as shown in Figure 5-10.

3. Tune peak anode cavity:

    a. Place a short circuit in place of the carrier cavity output loading iris. Use the normal coupling iris between the $\lambda/4$ waveguide section and the peak cavity (henceforth called "carrier interstage iris"). Use an iris with twice this value of coupling at the peak cavity output.

    b. Tune the anode cavity to resonance at the operating frequency (drive to the carrier cavity should be disabled for this test). This procedure trims anode cavity tuning to compensate for coupling iris reactive effects.

4. Tune carrier anode cavity:

    a. Place a short circuit in place of the peak cavity to $\lambda/4$ waveguide section iris (henceforth referred to as "peak interstage iris"). Use the normal carrier interstage iris.

    b. With very slight rf drive applied to the carrier input, tune the anode cavity to resonance using the rf sampler output for peak cavity rf voltage indication.

5. Restore all irises to their normal value for Doherty operation (per 3-a and 4-a).

6. The input cavities for both amplifiers should be set for matched input at full peak output ratings for initial operation of the Doherty amplifier.

7. Apply normal "tune-up" plate voltage (about 1 KV); apply normal Class B bias to the carrier stage and greater than two-times cut-off bias on the peak stage.

8. Apply rf drive to the carrier stage; gradually increase until approximately normal grid current is obtained. The following conditions should be observed:

   a. Lighter than normal carrier anode cavity loading (ideally half of normal).

   b. About one-half normal power output.

   c. The voltage sample in the peak cavity should lag that of the carrier cavity by 90°.

9. Reduce peak bias to the design value for the anode voltage used and apply approximately the same drive level as applied to the carrier stage in step 8. Slight plate current should be noted. If necessary, some adjustment of peak bias may be made at this point to adjust for slight plate current.

10. Increase peak drive (or reduce bias). Total rf output and carrier loading should both increase as predicted for Doherty operation. (Trim signal phase at the carrier input for 90° lead relative to the peak input phase if necessary.)

11. Some "cut and try" will be necessary at this point to balance cavity loadings, input phasing and relative drive levels, input tuning and matching and biasing. During this phase it will probably be found that varying bias with drive level may be necessary in order to obtain linear output versus total input drive. Record this data for later use in setting up the dynamic bias characteristic as required.

12. In the final system, incorporate dynamic bias into circuit and trim carrier and peak bias characteristics for best TV waveform linearity. Alternately use input matching scheme of Figure 3-7(b) in lieu of dynamic bias. (This step will not be necessary if satisfactory linearity and efficiency characteristics are obtained with fixed bias and the test type input circuit of Figure 5-11.)

5.2.2-3 Test Results

The above tuning procedure was used successfully in the set-up of the Doherty circuit. Testing was carried out and with light loading at 1000 volts on the Y1498 anodes to avoid excessive coupling into the anode bypass capacitor peripheral resonance. (This coupling proved to be much tighter with the current asymmetry caused by the twice normal peak output coupling; see step 3-c above.) The test results showed marked improvement in efficiency compared to a Class B amplifier with the same per stage output power. (This condition results in the same anode loading and output-to-cavity-loss ratio for both cases.) Performance of typical

Doherty operation is compared to average performance of the single cavity amplifier operating at one-half the output of the Doherty amplifier, using performance of the two cavity amplifiers in the single Y1498 amplifier tests reported previously. Typical results of this test were shown previously in the results of Section 3.2.2-3.

TV signal performance is indicated in stairstep waveforms presented earlier in Figure 3-14. Some sync stretching and differential gain/phase correction would be required as is normal television broadcast practice. The requirement for the use of dynamic bias is not strongly indicated at this time due to the good results indicated in these waveform measurements. This aspect requires further evaluation.

5.2.2-4 Dynamic Bias Circuit

The dynamic grid bias circuit is intended to vary the grid bias as a function of the instantaneous rf drive level. This is desirable for several reasons with the Doherty amplifier. The drive on the carrier tube causes saturation at carrier level, and badly overdrives the tube for higher input levels. Also, as the drive is increased, the drive on each grid does not increase linearly due to changing input impedance levels as the grids draw increasing current. By varying the bias in a manner similar to that in Figure 3-9, these non-linearities can be eliminated. In addition, efficiency will be increased if the carrier stage bias changes operation to the Class C region at high signal levels.

The dynamic grid bias circuit diagram was shown in Figure 3-8 and is repeated here for reference. Due to the high grid current possible, the circuit requires a low output impedance. A complementary PNP-NPN transistor pair operating push pull in an emitter follow configuration was selected. The two transistors chosen,

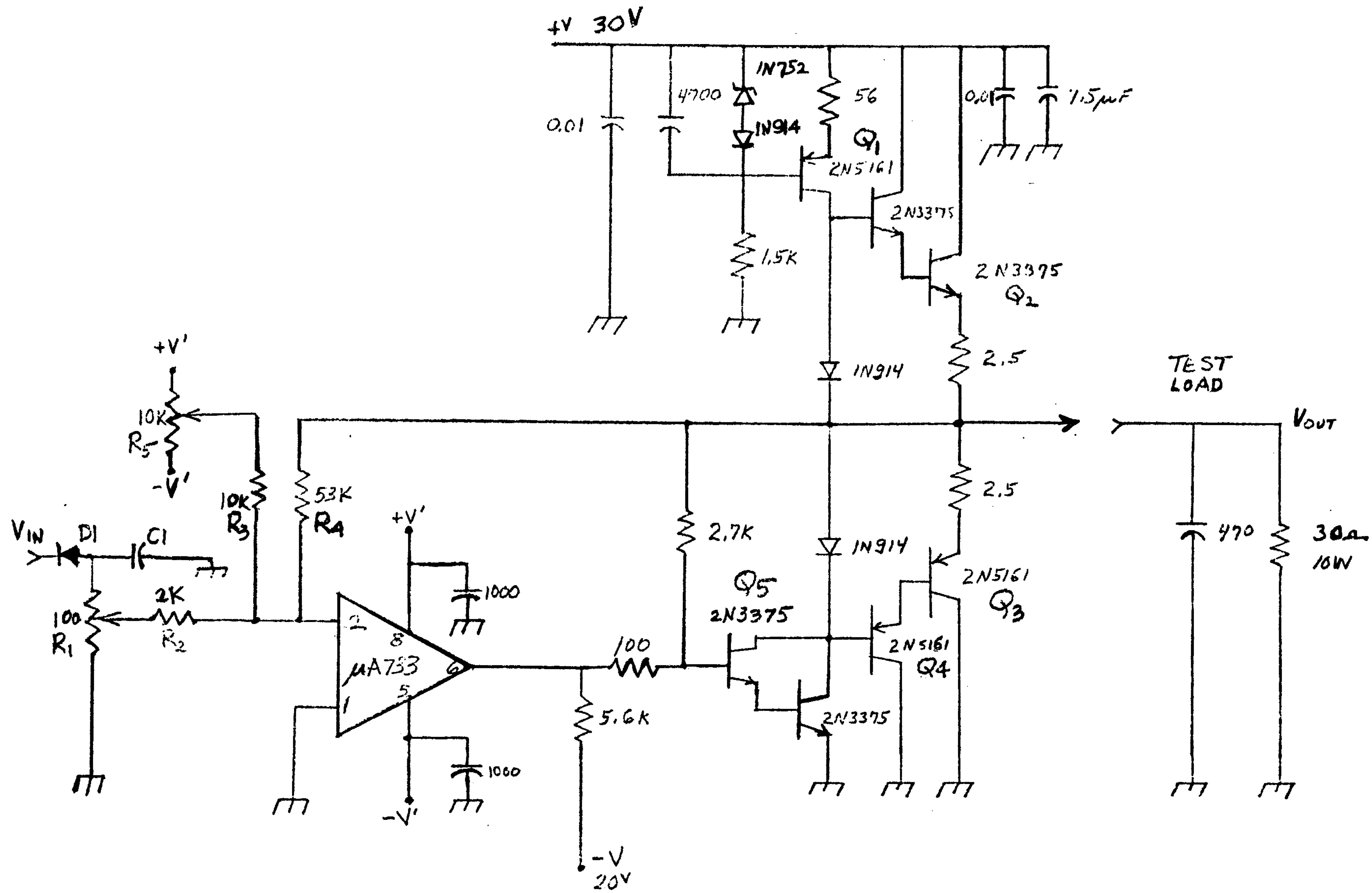


FIGURE 3-8. DYNAMIC BIAS SCHEMATIC FOR DOHERTY AMPLIFIER

Q2 and Q3, are assumed to have worst Betas of 5. The Q_1 and related circuitry is a constant current source set to deliver 100 ma which either pulls Q2 up with enough base drive, or is bled through the two IN914 diodes and through Q_4 and Q_3 to ground.

The "dead band" typical of the push-pull configuration is eliminated by allowing a small quiescent collector current to flow in Q_2 and Q_3. This current is set by the two 2.5 ohm fixed resistors between the two transistors.

The fact that the emitters of Q_2 and Q_3 can swing over a voltage range of about 40V requires the constant current source to supply the proper drive to Q_2. The dissipation of Q_4 is sized to handle the 100 ma from the current source plus the base current of Q_3 required to pull Q_3 down.

Q_5 is used as a low power transistor in the Darlington hookup to provide the necessary current gain between the μa733, which has a maximum sink current of only 2.5 ma, and the output of Q_4, which has a Beta of only 5.

The rf input is envelope detected by D1, C1 and R1 with the gain set by R1. The breakpoint of the detector is set at 100 MHz which is far enough away from the video band to cause no appreciable attenuation. Potentiometer R_5 is used to correct the output offset of the μa733 or to provide a fixed bias on the output.

Overall gain is set by feeding back the output to the input summing junction through R4. Initially, the gain is set at unity but can be easily changed by changing R4. Overall phase shift through the amplifier is minimized by keeping the closed loop gain low and using UHF transistors where necessary.

Preliminary tests have been run which indicated two problems, neither of which presented any great difficulty in their solution. The μa733 integrated circuit has a small output voltage swing capability in this application due to its particular differential output configuration. The Q_5-Q_4-Q_3 amplifier chain was designed to insure an adequate gain following the μa733. Secondly, the phase shift of the 2N3375 was measured to be 15° at 5 MHz, and the overall circuit phase shift was much greater. The proposed solution is to use a phase correction network in the circuit to compensate for this shift.

The breadboard test circuit used in evaluation of the circuit was shown in Figure 3-31.

5.3 AURAL CHANNEL AMPLIFIER - TASK 3

5.3.1 Specifications

The aural amplifier has been designed, fabricated, and successfully tested at the 500 watt level as required. Specifications were generated in the Task 1 System Study and indicated in Appendix A-3; the following are supplementary specifications for this amplifier:

Electrical

Gain	20 dB (objective)
Load Characteristics	1.3 VSWR (max.)
Tube type	GE Y1498
Plate Voltage Supply	1500 volts (typical)
Test Points	Monitoring points for all significant currents and voltages including rf input and output cavity voltage will be provided.

Other electrical objectives were:

- High efficiency and minimization of rf circuit losses
- Design adaptability for later space use
- Avoidance of multipactor discharge phenomena

Thermal

Anode cooling	Water, 0.25 gpm (nominal)
Cavity Parts and Other Tube Parts	Conduction or radiation (no forced air)
Heat Sink Temperature	100°C (max.) for conduction cooled surfaces 60°C (max.) for water inlet temp.
ΔT between adjacent tube seals	100°C maximum
Maximum tube seal temperature	300°C maximum

Additional design goals were:

- Minimize electrical insulators in series with heat flow path
- Anode at same electrical potential as spacecraft heat rejection system

Mechanical

Cavity Construction	Breadboard design should be adaptable to space-type hardware with minimal changes - the approach for doing this should be defined in the task final report. Design features should include: Ruggedness Avoidance of excessive mechanical stresses on the tube and other amplifier components due to external forces encountered during normal use including thermal expansion. Avoidance of excessive weight. Cavity should be readily dismantled for tube replacement, developmental changes, etc. Avoidance of excessive thermal detuning effects (or provide for later incorporation of this feature).
Auxiliary Circuit Construction	Package bias and similar circuitry in a neat fashion. From the standpoint of personnel protection, this circuitry may be mounted as a subassembly in the test power supply rack. However, it must be designed so that it can readily be removed for delivery to the customer upon completion of the contract.
RF Connectors	
Input	Type N Coaxial
Output	Half height WR975 waveguide
Power Connectors RF Test Point	BNC, Female
Electrical Hazard Avoidance	No exposed voltages greater than 24 volts rms above reference ground. Connector design should permit ease in removing cavity from test setup and dismantling of the cavity.
Compatibility with Breadboard Circuit	Designs should be periodically reviewed with the project engineer to assure compatibility with all electrical and mechanical interfaces in the breadboard circuit.

Other mechanical goals were:

- No stresses due to temperature differentials between tube electrodes
- Use of materials which are lightweight, give long life operation in space, and have minimum outgassing characteristics.
- East of adjustment of coupling and tuning
- Ability to alter any circuit element without scrapping costly parts or assemblies.

Personnel Safety

High Voltage	All terminals more than 24 vrms above ground will be adequately insulated or shielded to prevent accidental contact by personnel.
RF Radiation	The level of all electromagnetic fields will be maintained below 10 milliwatts per square centimeter at all points accessible to personnel.
Hot Spot Temperature	All points on the outside surfaces of the circuitry which operate at temperatures above 100° C will be adequately shielded to prohibit personnel contact insofar as practicable.

5.3.2 Selected Approach

The aural high power amplifier configuration selected has an rf grounded grid with a dc grounded anode. The grounded anode is advantageous in permitting the use of a high temperature, high flux density, heat pipe for anode cooling without a high voltage insulator between anode and heat pipe. Such an insulator would make a larger heat radiator necessary since the significant temperature drop across the insulator could be countered only by a lower temperature (and, therefore, larger) heat radiating plate. Thermal isolation was provided between anode and the cavity, however, to allow the grid to remain moderately cool and to minimize thermal detuning of the cavity.

Consideration of the circuit type also included the potential high power breakdown conditions that might arise in the high vacuum of space. That is, multipactor

breakdown in the anode cavity can be avoided by biasing one side of the output cavity relative to the other, at least in the regions where rf voltages provide a possible breakdown condition. A proposed method of insuring no breakdown is the inclusion of a disc-shaped biased electrode near the anode of the tube within the cavity.

The tube selected for the Doherty amplifier is also considered to be the best for the aural stage. The Y2042, which is temporarily replaced by the Y1498, will provide the best performance of any tubes in the performance range required.

5.3.3 Cavity and Circuit Design

The preferred circuit applicable to planar triodes like the Y1498 tube in the UHF region of the spectrum is an rf-grounded-grid circuit employing coaxial resonators. This circuit was presented in simplified form in Figure 3-16, and a detailed cross-section of the cavities is in Figure 5-12. A dc grounded anode is used as was noted above.

The tube inter-electrode capacitances load the resonant transmission line sections, requiring that the cavity lines be shorter physically than $\lambda/4$ (or $3\lambda/4$). The line X_c is equal to $-Z_o \tan \beta \ell$, where β is the phase constant in radians per unit length. In the present case, with a Y1498 tube and f_o=829.75 MHz, the foreshortening of the cavity necessitated by the grid plate capacitance and the use of a high Z_o line reduces the length of the output cavity line to a dimension less than the diameter, making the line, in effect, a flat "pill box" shape, better described as a radial transmission line cavity. Similarly, the grid-cathode capacitance is of such magnitude as to make the design of a $\lambda/4$ grid-cathode line impractical, so a $3\lambda/4$ coaxial cavity is used for the cathode input circuit. The latter is folded to reduce physical size.

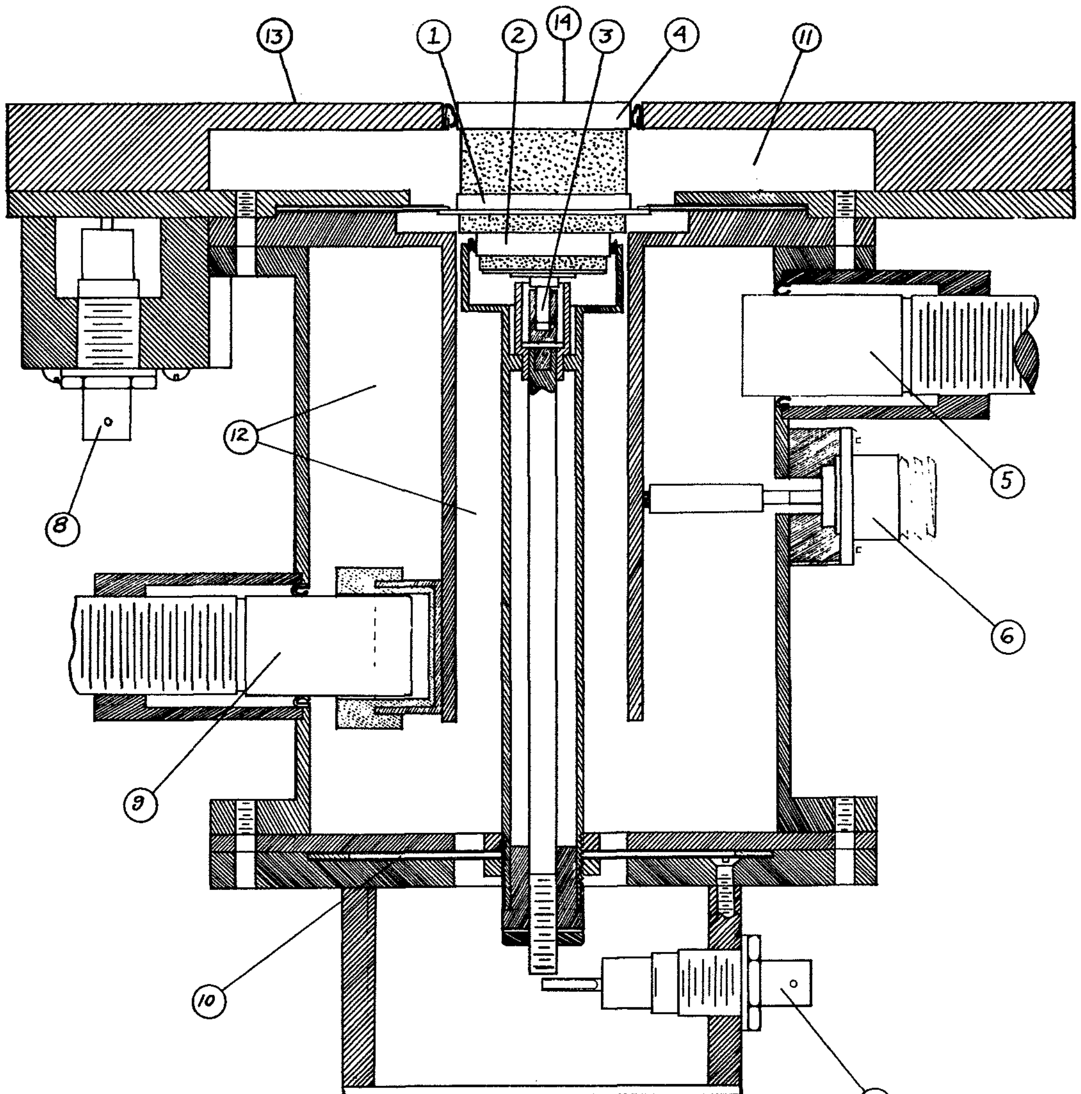


Figure 5-12. Cross Section of Aural Channel Amplifier

KEY

ITEM	DESCRIPTION
1	GRID CONTACT SURFACE
2	CATHODE CONTACT SURFACE
3	HEATER CONTACT PIN
4	ANODE CONTACT SURFACE
5	INPUT COUPLING ADJUSTMENT PLUNGER
6	RF INPUT CONNECTOR
7	HEATER AND CATHODE POWER CONNECTOR
8	GRID BIAS CONNECTOR
9	INPUT TUNING PLUNGER
10	CATHODE BYPASS CAPACITOR
11	ANODE-GRID CAVITY RESONATOR
12	CATHODE-GRID CAVITY RESONATOR
13	MOUNTING SURFACE AND CAVITY HEAT SINK INTERFACE
14	TUBE ANODE HEAT SINK INTERFACE

The simplified schematic of Figure 3-16 showed the necessary dc blocking capacitors as used. The most convenient location for low-loss rf capacitors in series with the electrodes was at the flat surface in the plane of the grid and at the short-circuited end of the folded grid-cathode line. The inner section of the amplifier cavity cross-sectional drawing, Figure 5-12, shows the hardware realization of the bypass capacitors in the circuit. Referring to this figure, a variable piston type capacitor (9) has been incorporated in shunt with the cathode line at a high impedance point for tuning. A second movable plunger (5) intercepts a portion of the rf magnetic flux which links the input coupling loop, thereby providing a means of adjusting the coupling.

Tests on the amplifier were conducted using a coupling loop to extract power from the anode cavity. In a final system, an iris in the wall of the cavity would couple the anode cavity directly to the output waveguide.

5.3.4 Mechanical Design

Certain features in Figure 5-12 may be noted with respect to the design objectives:

1. The tube anode is dc grounded and thermally isolated from the cavity structure permitting the cavity to operate at a lower temperature than the anode.
2. Some flexure is premitted between the tube and the eavities so as to relieve stresses that might arise due to separate mountings for the cavity and the heat transport system attached to the anode.
3. The upper surface of the anode cavity provides a convenient mounting surface and a heat transfer interface.
4. The copper flange soldered directly to the grid contact surface provides a thermal path for grid cooling.
5. The cavity is easily dissassemblied for examination and modification. The breakdown is such that individual elements may be changed without scrapping associated parts.

6. No dangerous voltages are accessable when the cavity is assembled.

7. The overall structure is compact and rigid.

5.3.5 Anode-Grid Test Cavity for Y1498 Tube

In order to determine the dissipative loading effect of the tube on the tank circuit, and to investigate the contact properties of titanium electrodes, a "cold" cavity was fabricated as a grid-plate tank circuit for testing. No dc blocking capacitors were incorporated, since it was not intended to apply dc power to the tube. The cavity resonance occurred at approximately 1 GHz with the first two Y1498 tubes received. The Q and transmission coefficient were found to be erratic; when the tube's anode ring was polished, the Q increased from about 120 to 300. The cavity parts were then plated with 0.5 mil of silver and covered with 25×10^{-6} inch of gold. The tube's electrodes were plated with about 0.5 mil of gold over a nickel strike. The Q of the cavity-tube combination was then about 800. The vacuum envelope of a plated electrically defective tube was intentionally pierced to observe the resultant effect on the Q of the cavity-tube combination; there was no noticeable change in Q. The test cavity was then heated to determine drift; from 25°C, to 125°C, its resonant frequency increased by about 3.5 MHz, which will be of no concern for this amplifier.

5.3.6 Test Results on Amplifier

Initial tests on the aural amplifier revealed a seriously low isolation between the input and output cavities. This was corrected by placing a quarter wave slot shorting section in series with the grid bypass cover (cathode side) to prevent the rf from leaking between the two cavities via the bypass capacitor. The result was 40 dB of isolation, which is sufficient for this amplifier. This modification interfered somewhat with the input cavity loading adjustment (part 5 in Figure 5-12 so subsequent adjustments were made using an external tuner.

Efficiency was measured to be about 44% which was considerably below the expected performance. After evaluating the possibilities and recognizing a similar difficulty in the Doherty stages, a thermal measurement to develop a heating pattern for the cavity in the vicinity of the bypass capacitor, and the result indicated the presence of a circumferential resonance mode. Thus, some modification is called for in subsequent amplifier designs. A substantial improvement of efficiency, to nearly 70% maximum, was observed when the amplifier was tuned to a significantly different frequency. Thus, a small amount of bypass capacitor redesign is required for high efficiency from the aural stage, but the amplifier was able to provide the required 500 watts output, even with this limitation.

5.3.7 Space Design Factors

Some concern exists about possible rf breakdown in the present design if used under space conditions. The input cavity should be free from multipacting; however, the output cavity dimensions are such that occurrence of multipactor breakdown is possible. The use of a biased electrode in the anode cavity is proposed as a means of suppressing any such tendencies. This electrode would be disc shaped, parallel to the anode surface of the tube, and could be connected to the plate supply potential.

Ionizing breakdown (dc and/or rf) could also be a problem due to the possible entrapment of air in the stacked disc mica capacitor assemblies used as grid and cathode rf bypasses. Either a proper bakeout procedure or the use of void-free capacitor assemblies may be a suitable solution to this potential problem. The latter approach may require development of effective rf bypass capacitor assemblies, probably involving metallizing of the mica surfaces.

5.4 RF COMPONENTS - TASK 4

5.4.1 RF Components and Transmission Lines

In determining a preferred type of RF component and transmission line, considerations included 3-1/8" coaxial line, half height WR975 waveguide, and possible ridged waveguides. The latter save on weight and size, but tend to be very low in impedance. Thus, the impedance matching sections required as well as the higher losses would make this approach more costly, and ridged guide appears to give little or no system improvement, especially in filters which might become quite lossy. It was not considered practical to use the ridged guide in this program, and the standard guides were selected for the breadboard circuit. Evaluation of low impedance line sections for vacuum breakdown susctptibility is included in Task 7, High Power RF Component Environmental Test Plan, described in Sections 3.7 and 5.7. Results of earlier analysis (Reference 2) were scaled for operating conditions and the results are summarized in Table 5-5 . Low multipactor susceptibility, low heating, and low insertion loss indicate the waveguide components are generally superior to coaxial line. Waveguide is bulkier than coax, but is comparable in weight and may be lighter when the required thermal control measures are applied to the coaxial line inner conductor. Both coaxial line and waveguide reactive harmonic filters appear to be susceptible to multipacting; this can be circumvented with leaky wall type absorptive filters (which are heavier and bulkier) or with closer spaced, low impedance versions. Insertion loss for the waveguide circuit is expected to be about 1/2 that of a coaxial version. On the basis of these considerations, half-height WR975 waveguide components were chosen for the breadboard circuit.

TABLE 5-5

COMPARISON OF RF COMPONENTS AT 5 KW

COMPONENT	OPERATING VOLTAGE KV	MULTIPACTING RANGE KV	HOT SPOT ΔT ^{o}C	COMMENTS
1/2 HEIGHT WR975	1.5	7.5 - 30.	0.5	
3-1/8" COAX	0.5	0.55 - 5.	10.	ONE STUB PER FOOT
WR975 HARMONIC FILTER	0.57	0.25 - 1.3	1.3	
COAX HARMONIC FILTER	0.5	0.18 - 2.4	75.	STUB AT ENDS (1 FT. LONG)
WR975 AURAL FILTER	3.4	7.5 - 25.	1.5	
COAX AURAL FILTER	8.4	13. - 30.	44.	
WR975 HYBRID	1.5	7.5 - 30	1.	ESTIMATED
COAX SLAB HYBRID	0.45	0.18 - 2.4	37.	STUBS AT CONNECTORS
COAX BRANCH HYBRID	0.41	0.4 - 3.0	20.	ESTIMATED; STUBS AT CONNECTORS

5.4.2 High Power RF Components

The completed assembly was shown in the photograph of Figure 3-18. After assembly, additional tuning was performed to improve VSWR in the passband. This resulted as:

	Before Tuning	After Tuning	
VSWR			
Visual Channel	1.26 to 1.30	1.10	
Aural Channel	1.15	1.10	
Insertion Loss	0.1 dB (Est.)		
	Length	Width	Height
Dimensions (Excl. Flange Hts.)			
Filter/coupler assembly	23.9 in.	20.0 in.	2.7 in.
To top of cavity	---	---	17.3
To top of couplers	---	---	5.1
Hybrid	24.0	20.0	2.7

Dimensions and weights of these units for spacecraft use can be reduced. The mechanical techniques used in these units are basically those of ground type equipment, which is an acceptable approach for an experimental breadboard transmitter.

5.4.2-1 Color Notch Filter Design

The color subcarrier image notch filter is a single-tuned circuit design, following current television design practices and the recommendations of previous studies (Sec. 5.3.2 of Reference 2). Since the width of the dual waveguide assembly is already about 20 inches, it was decided to use a top-coupled cavity, which could be configured to lie parallel to the waveguide between visual amplifier and load in a spacecraft design. This arrangement would result in a filter assembly height of less than 6 inches (including the waveguide to which it is coupled) and a length of about 20 inches. Several configurations, including the basic "T" top wall coupled configuration fabri-

cated for the transmitter breadboard are indicated in Figure 5-13 . The other four configurations show three folded versions, any one of which could be used for a flight system.

In addition to the specifications of Appendix A-4, the following supplementary specifications apply:

- Electrical:

Insertion Loss for the visual channel	0.10 dB (maximum allowable) at 825.25 MHz and averaged over 824.0 MHz to 829.5 MHz, more than 20 dB at 821.67 MHz

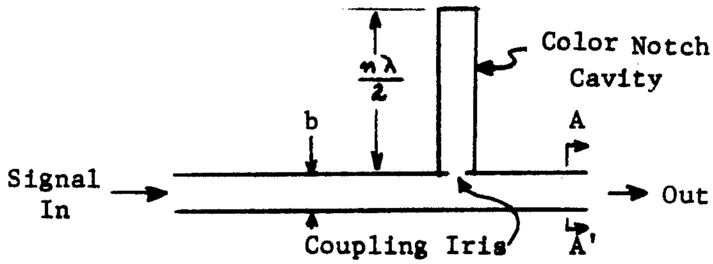
- Mechanical

General Arrangement	Attach to upper broad wall of guide
Length	Minimize, 28 inches maximum axial length (24 inches, typical)
Flanges	One-half height WR975 flat face, tolerances should permit application of flange-to-flange connections without necessity for auxiliary rf "gasketing" between flange faces.
Material	Waveguide and most of structure to be aluminum alloy construction.
Finish	Protective film of aluminum parts as described in MIL-C-5541, silver plate cuprous alloy materials, other materials per good commercial practice.
Cooling	Convection, ambient air.

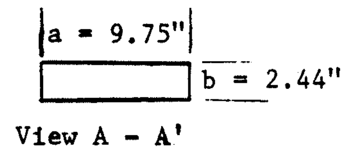
- Environment:

Ambient Temperature	18°C to 35°C
Relative Humidity	Less than 80%

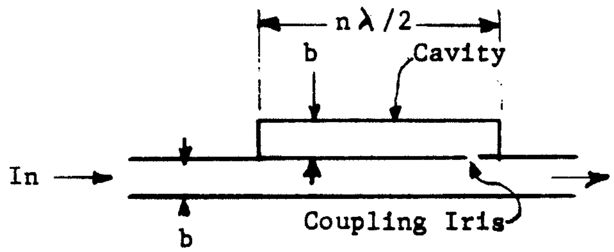
The waveguide assembly will contain a dual directional coupler in each channel to allow monitoring of power output, reflected power, and load VSWR. The color image filter is located between the output cavity of the final amplifier stage and these monitoring directional couplers; its performance is included in the overall performance of the visual amplifier chain.



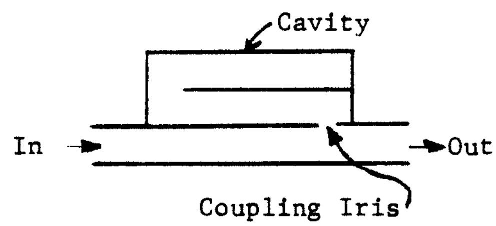
(a) Basic Top Wall Coupled Notch Filter



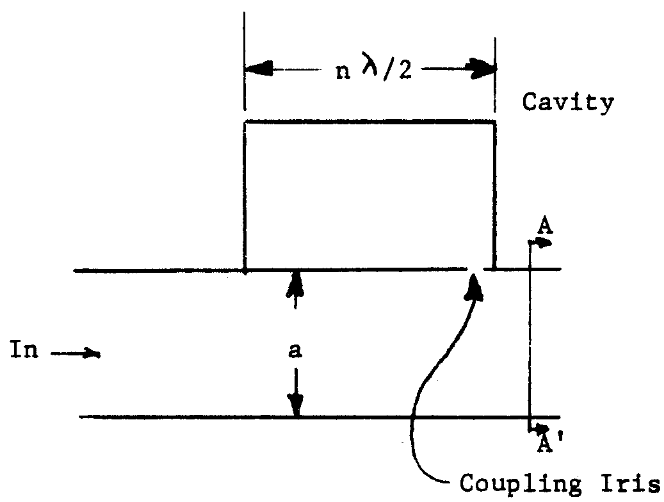
(b) ½ Height WR 975 Waveguide



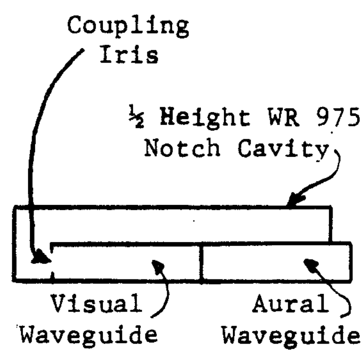
(c) "Minimum Height" Configuration of Notch Filter



(d) Folded Version of Notch Filter



(e) Side-Wall Coupled Notch Filter



(f) Side-Wall Coupled Notch Filter with Filter Cavity Folded Over Dual Waveguide Run (Cavity is coupled to visual waveguide only. Head on view of visual and aural waveguides shown as in (b) cross-section view).

FIGURE 5-13. NOTCH CAVITY CONFIGURATIONS

The basic circuit for the filter is a single tuned parallel-resonant circuit in series with the waveguide run, as shown in Figure 5-14. Values of parallel resonant circuit L and C are such that loaded Q is very high, and the circuit provides a high attenuation at the color subcarrier image frequency. It becomes a very low impedance at frequencies within the visual transmitter passband and offers little loss to these frequencies. A simple analysis based on the filter circuit of Figure 5-14 was used to determine the filter design characteristics.

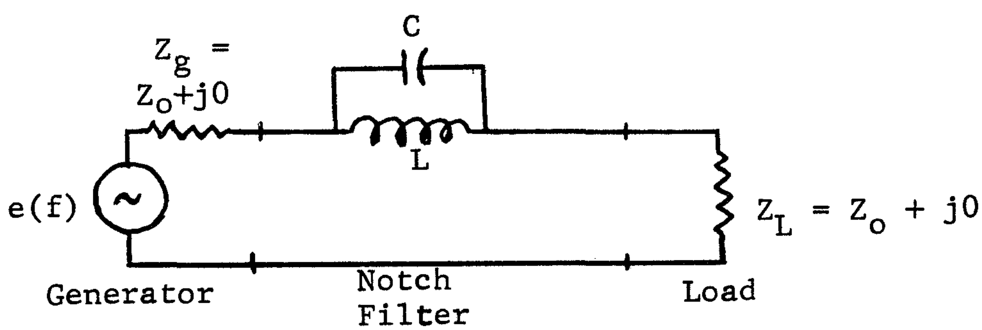


Figure 5-14. Notch Cavity Circuit

The analysis indicated an unloaded Q of 18,200 is desirable to obtain a 20 dB rejection at 821.67 MHz but a loss of 0.1 dB at the 825.25 MHz carrier frequency. The Q is attainable with a half-height WR975 dimensioned cavity. Details on the design of iris couplings and band reject cavity filters are given in Section 5.10 and Chapters 8 and 12 of Reference 9. The cavity length is slightly less than $\lambda g/2$, this providing a small capacitive reactance to balance the iris inductive component. Selection of the appropriate relations for the particular case in a given requirement can be derived from information given in this excellent reference.

In a spacecraft version considerable improvement in thermal characteristics could be provided if the notch filter is mounted against the broad wall of the visual waveguide as in Figure 5-13. This also gives rise to a weight reduction since the waveguide and cavity can now share a common wall. Additional weight could be saved by machining away large portions of the waveguide wall thickness leaving a thin-skinned, ribbed member as required for rf, structural, and thermal conduction characteristics. The lower broad wall surface will be attached to a heat sink for thermal control in the spacecraft version.

Measured performance of the filter for TV channel 73 is:

Notch attenuation	20 to 23 dB
VSWR at visual carrier	1.23
Loss at Visual carrier (calculated from VSWR)	0.1 dB

A tuner is required in the overall assembly to achieve an acceptable VSWR; the introductory paragraph of this section indicated a 1.10 VSWR with tuning.

5.4.2-2 Directional Couplers

The reflectometer-type directional coupler has adequate performance in the UHF television band and was chosen for this application primarily on the basis of compactness. Two sets of directional couplers are included in the waveguide assembly of Figure 3-18, one set in each of the two RF channels. Like all the waveguide components, these are fabricated in half-height WR975 waveguide, and the pair have a common narrow wall. Each set of couplers includes a forward and a reverse coupler. The color notch filter precedes the coupler in the visual channel, and the aural amplifier feeds directly into the coupler in the aural channel; outputs feed into the 3 dB hybrid. Some of the pertinent specifications for the directional couplers are:

- Electrical

Coupling ratios	For the frequency range 824.0 MHz to 830.0 MHz
a. Forward Visual Coupler	-56 dB $\pm$ 0.2 dB
b. Reverse Visual Coupler	-46 dB $\mp$ 0.2 dB
c. Forward Aural Coupler	-36 dB $\pm$ 0.2 dB
Directivity (all Directional Couplers)	30 dB (minimum) for the 824.0 MHz to 830.0 MHz frequency range
Directional Coupler RF Output Fittings	Type N - Female
Operating Transmission Range	Above specifications shall apply for convection cooling with 18 to 35° ambient temperature and up to 5.0 kW peak synchronizing visual power and 0.5 kW CW aural power.

- Mechanical

General Arrangement	The directional couplers will be a part of the output waveguides from the two output amplifiers. No components are to be attached to the lower broad wall of the waveguide assembly.

An illustration of the basic form of the coupler was shown in Figure 3-19a, repeated here for convenience. A small loop is introduced into the waveguide; each end of

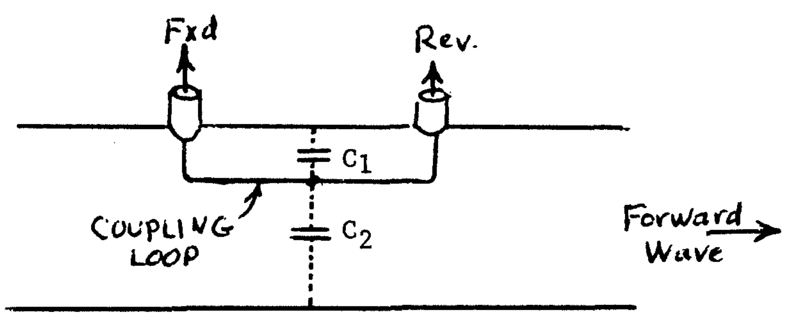


Figure 3-19a. Basic Reflectometer Type Directional Coupler

the loop is connected to loads (or external instrumentation) by means of coaxial lines of some convenient impedance Z_o. The loop area and orientation with respect to the waveguide axis controls coupling to the magnetic field within the waveguide. Coupling to the electric field within the waveguide is also made with the loop. By adjusting loop size and shape, with a particular load Z_o at either end, and with $C_1 \gg C_2$, the currents induced by electric and magnetic field can be made to add in one load and cancel in the other for waves propagating in a single direction in the waveguide. In the case shown, the Forward wave sample couples to the left hand coaxial post. For this condition of coupling, a wave traveling within the waveguide in the opposite or Reverse direction couples to the right hand port. With proper adjustment, a particular port will have over 30 dB ratio in its coupling to Forward and Reverse waves. This "Directivity" allows measurement of Forward and Reverse wave amplitudes, from which VSWR can be calculated for a particular value of coupling. The coupler can be used with a thermistor mount type microwave power meter to measure power in the wave. Figure 3-19b showed a more practical arrangement for fabrication. Here the forward and reverse monitors are separate units to provide better isolation of operation.

The couplers are included in the photograph of Figure 3-18. The coupler assemblies are mounted on the load side of the filter so that filter effects will be included in the signal output sample. A second set of couplers is mounted on the aural channel waveguide.

Measured performance of the coupler assembly for TV channel 73 is as follows:

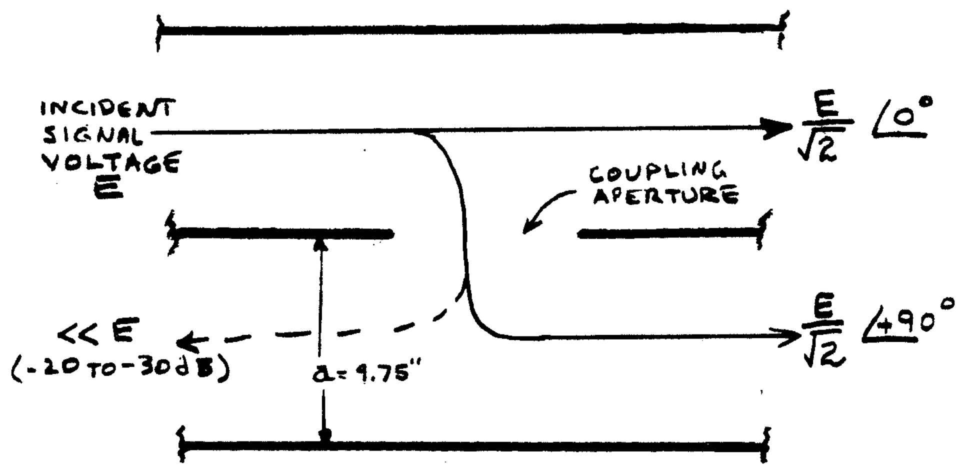
	Coupling dB	Directivity dB
Visual Forward	50.2	$>$ 30 dB
Visual Reverse	40.1	$>$ 30 dB
Aural Forward	40.2	$>$ 30 dB
Aural Reverse	40.2	$>$ 30 dB

5.4.2-3 3-dB Sidewall Hybrid

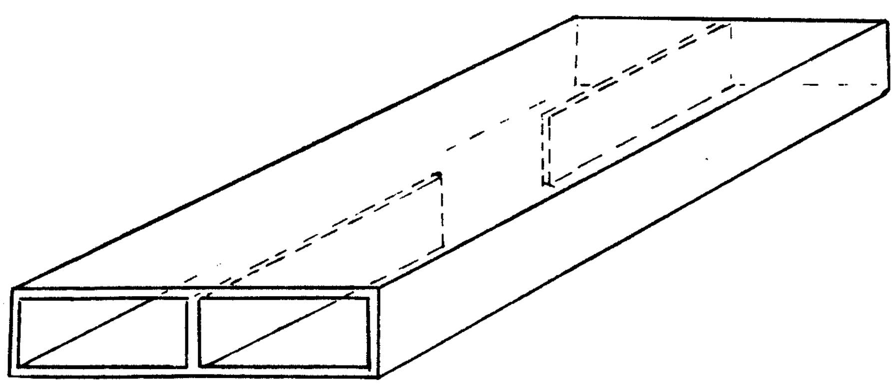
The hybrid is a 30 dB short slot type coupler fabricated in one half-height WR975 waveguide. The flanges used to couple the hybrid to mating components are of the dual flange type which permits the design of the hybrid to have shortest possible axial length. Detailed specifications for the hybrid were given in Appendix A-4; some supplementary specifications are:

Electrical	
Sliding load termination VSWR	1.03 (maximum)
Mechanical	
Waveguide	half height WR975
Length	Minimize, 24 inches maximum axial length
Flanges	Dual one half height WR975 common narrow wall, flat face, tolerances should permit applications of flange to flange connections without necessity for auxiliary rf "gasketing" between flange faces.
Materials	Waveguide and majority of structure to be aluminum alloy construction.
Finish	Protective film on aluminum parts as described in MIL-C-5541, silver plate cuprous alloy materials, other materials per good commercial practice.
Cooling	Convection, ambient air
Environment	
Ambient Temperature	$18^{o}C$ to $35^{o}C$
Relative Humidity	less than 80%

Detailed design of a sidewall coupler is described in the literature[10]. It consists of two parallel waveguides sharing a common wall which contains a coupling aperture. By properly dimensioning the aperture, 3dB coupling is obtained so that a wave entering either input port is divided equally and transmitted to two output ports. The other input port is isolated from the first and receives little power (typically -20 dB). Additional tuning elements are normally incorporated to give increased bandwidth and as means of optimizing performances of the hybrid. Figure 5-15 is a sketch of the hybrid.



a) Top View of Sidewall Hybrid



b) Basic Sidewall Hybrid Construction

Figure 5-15. 3-dB SIDEWALL HYBRID

Measured performance of the half-height WR975 sidewall hybrid is given below.

Coupling	-3.0 dB
Isolation	33 dB at 824.0 to 830.0 MHz
VSWR	1.03 at Band Center 1.07 at Band Edges

The hybrid is included in the assembly photographs of Figure 3-18.

5.4.2-4 Waveguide to Coax Transitions

The transition is a dual unit which is used to couple the two waveguide outputs of the 3 dB hybrid to their respective coaxial dummy loads. The dual waveguide section used has a special dual waveguide flange which permits mating with the 3 dB hybrid output. Specifications were indicated in Appendix A-4; supplementary ones are:

- Mechanical

General Arrangement	See Figure 3-18. No components should be attached to the lower broad wall of the waveguide assembly; use half-height WR975 waveguide.
Length	Minimize, 12 inches maximum axial length (10 inches, typical).
Flanges	Dual one-half height WR975 common narrow wall, flat face, tolerances should permit application of flange-to-flange connections without necessity for auxiliary rf "gasketing" between flange faces.
Finish	Protective film on aluminum parts as described in MIL-C-5541, silver plate cuprous alloy materials, other materials per good commercial practice.
2.7 Cooling	Convection, ambient air.

- Environment

3.1 Ambient Temperature	$18^{o}C$ to $35^{o}C$
3.2 Relative Humidity	Less than 80%

Each transition section contains a conventional T-bar coupling between half-height WR975 waveguide and a 1-5/8 inch coaxial lines. The intrinsic bandwidth of the transition is greater than that of a television channel. Measured performance is:

VSWR	1.03
Bandwidth	824.0 to 830.0 MHz

5.4.3 Vestigial Sideband Filter

The vestigial sideband filter is to be designed to shape the transmitted video rf signal in compliance with U.S. television standards as outlined in EIA Standard RS-240.(3) Additional requirements are for minimal size and weight consistent with moderate insertion loss. Lightweight is considered a significant feature, so the VSB filtering will be accomplished at a low power level, at a few watts where considerable loss can be tolerated with little impact on the efficiency of the overall transmitter. Its construction will utilize a TEM wave structure such as stripline or coaxial lines. Basic specifications for the VSB filter were given in Appendix A-4; supplimentary ones are:

Electrical	
Operating Frequency	TV Channel 73 (824 to 830 MHz)
Bandpass Characteristics	Per Figure 3-20
Input Power	10 Watts (max.)
Insertion Loss	3.0 dB (max.) from (f_o - 0.75) MHz to (f_o + 4.2) MHz
VSWR	1.5:1 (max.) from (f_o - 0.75) MHz to (f_o + 4.2) MHz
Mechanical	
Cooling	Design for conduction cooling of all elements
Construction	Design should be readily adaptable to rugged, lightweight construction for spacecraft use. Minimum size is also desirable.
RF Connectors	Coaxial, Omni Spectra OSM or Equivalent
Compatibility with Breadboard Circuit	Designs should be periodically reviewed with the project engineer to assume compatibility with all electrical and mechanical interfaces in the breadboard circuit.
Environment	
Ambient Temperature	18°C to 35°C
Relative Humidity	Less than 80%

The filter design makes use of the phase sensitive properties of a 3 dB quadrature hybrid as was shown in the block diagram of Figure 3-21. This filter will give sharp skirts without the use of a large number of resonant cavities.

In the two quadrature hybrid with the two tuned circuit terminations, the magnitudes of the reflections are unity; if the relative phases of the reflections are in phase, the hybrid will pass all of the incident energy, but if the phases of the reflections are 180° relative to each other, all the energy is reflected back into the source.

The design of this filter is based on Reference 5 and uses several equations from that document. Filter voltage response is given by the equation,

$$R = \frac{2j}{k_3 + k_4}$$

where k_3 and k_4 are the reflection coefficients at ports 3 and 4 of the hybrid network.

The reflection coefficients magnitudes and phases can be determined for various terminations, and this information can be used to determine k_3 and k_4, and thus the overall filter response. Single and double tuned cavities are of specific interest. An analysis showed that as long as the Q's of the terminations are large, the filter is not dependent on Q.

The simplest filter would utilize single tuned cavities as reactive terminations. However, the bandwidth of such a filter is not sufficient to give the required 20 dB of attenuation over the 4.5 MHz bandwidth. In order to investigate this filter for feasibility, the phase variations of the single cavity termination was determined for various values of coupling, T. Then a duplicate transparency overlaid the first series of plots. By sliding the graphs relative to each other and visually comparing the curves, the filter characteristic can be readily recognized. Maximum attenuation occurs for maximum phase difference for the curves. The phase response for the single cavity is given in Figure 5-16. This method of analysis showed that two single cavity reactive terminations are not sufficient to give the required 20 dB of attenuation over a 4.5 MHz bandwidth.

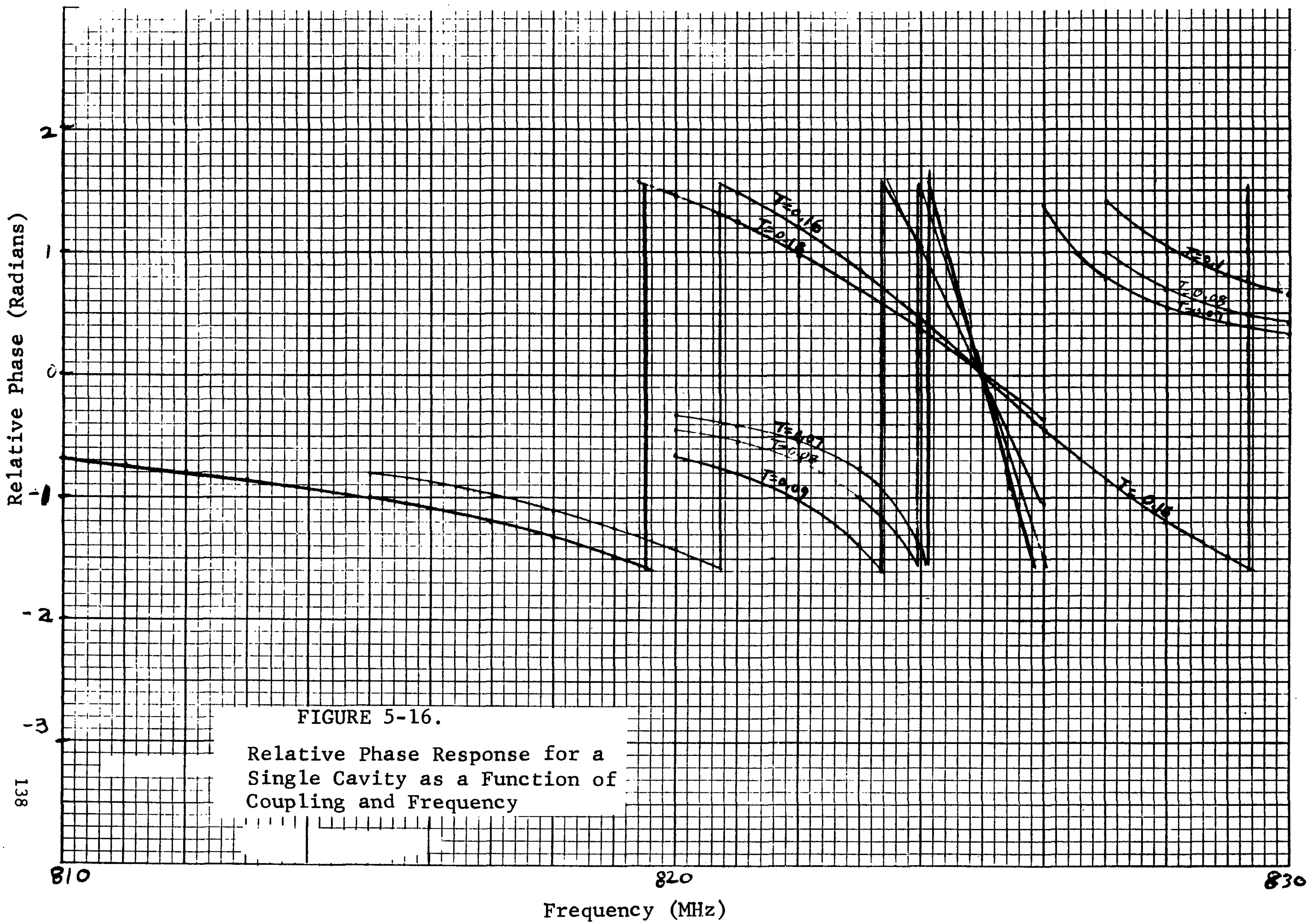


FIGURE 5-16.

Relative Phase Response for a Single Cavity as a Function of Coupling and Frequency

The next filter considered used a single tuned cavity at one port of the hybrid and the double tuned cavity at the second port as was shown in Figure 3-21. The phase response of double tuned cavity was plotted for various values of coupling. The results were given in Figure 3-22. This configuration should give the required response.

An experimental filter was fabricated, as was shown in Figure 3-23, to verify theoretical conclusions. This filter was fabricated in a strip transmission line form with overall dimensions of 6 x 10 x 3/4 inches. The ground planes were fabricated from 1/8 inch aluminum plate and the filter structure was made from one mil brass shim stock. Four 1/8 inch polystyrene plates were sandwiched to fill the volume between the ground planes and to support the filter structure. Polystyrene has similar electrical characteristics to PPO material, which a flight model filter would use. The filter structure was made from one mil brass shim stock except for the resonators which are silver foil.

Initial tests on the filter showed the rejection band to be broader than predicted and the filter skirts to be less stepp than predicted. A swept frequency measurement of the transmission response is shown in Figure 5-17 . The resulting response for the visual carrier placed on the right hand side of the curve is -3 dB at the -0.75 MHz point, and -20 dB at the -2.75 MHz point (rather than at -1.25 MHz per the specificat-ion). The -20 dB rejection bandwidth is 2.5 MHz wide, about 0.5 MHz less than desired. An alternate adjustement resulted in -20 dB rejection at 4.0 MHz from the -0.75 MHz point, and a -20 dB rejection bandwidth of 7.1 MHz as shown in Figure 5-18.

Correction derived from the test results indicate that resonator Q is lower than necessary to meet specified performance. Performance objectives should be achieved by modifications which add one resonator, increase spacing between conductor and ground plane to at least one inch, and remove much of the dielectric material adjacent to

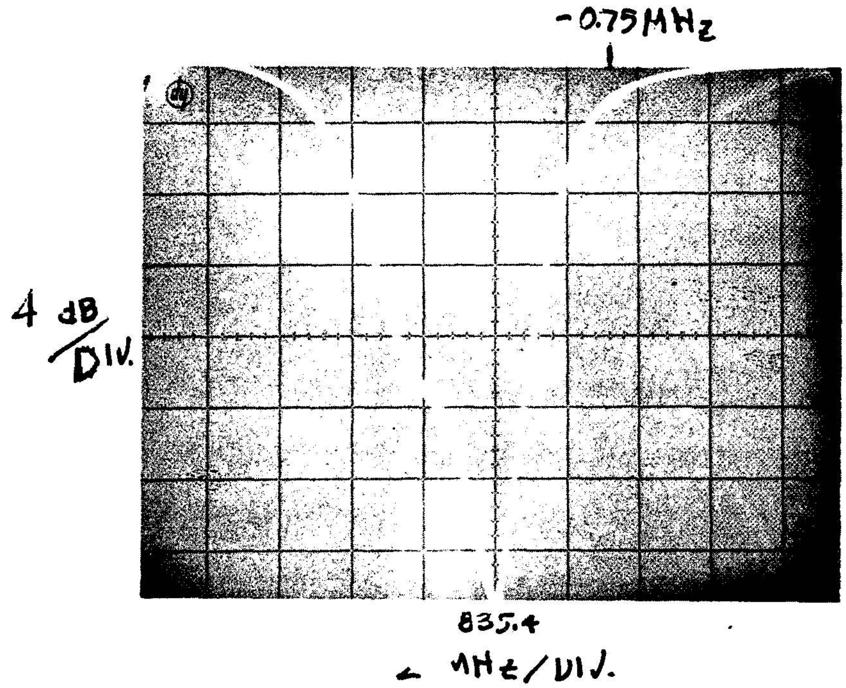


FIGURE 5-17. SWEPT FREQUENCY RESPONSE OF VESTIGIAL SIDEBAND FILTER

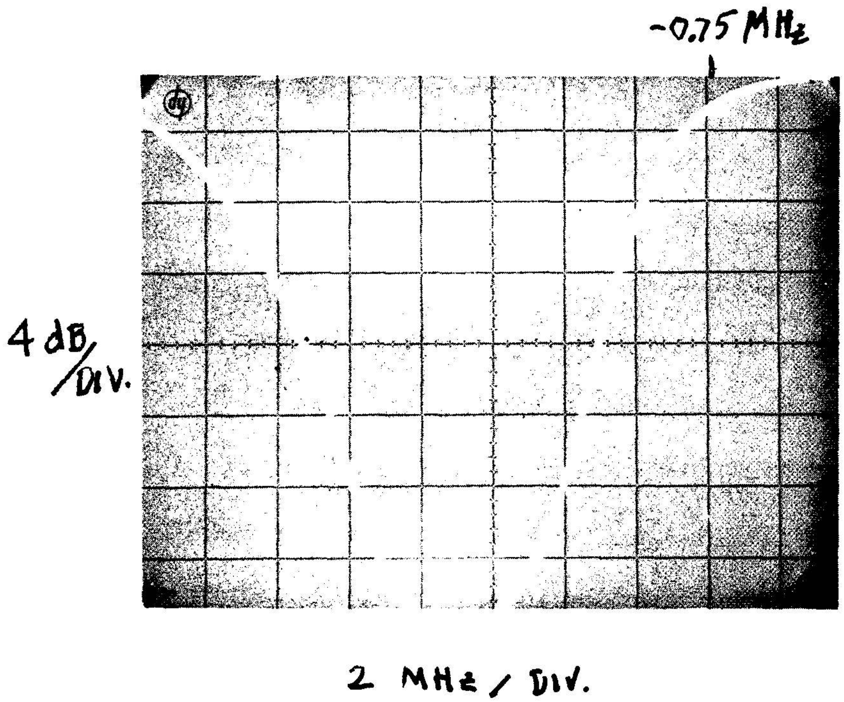


FIGURE 5-18. READJUSTED RESPONSE OF VSB FILTER

the resonators. Consideration should also be given to alternate construction with one-inch coaxial line cavities (or the next larger standard size of 1-5/8 inch OD) where adequate resonator Q will be obtained.

5.5 MONITOR AND PROTECTIVE CIRCUITS - TASK 5

5.5.1 System Aspects

The purposes of the monitor and protective circuitry are to sense electrical faults that may occur (dc and rf), fire a protective crowbar (dc), and disconnect the appropriate source(s) (dc or rf). Emphasis in this task is on the dc crowbar spark-gap since this is the critical component in the circuit.

The crowbar circuit diverts the current of the power supply and its filter elements from the high power tubes, in the event of an internal tube arc, to ground. This prevents excessive energy dissipation in the tube arc and subsequent damage to delicate tube elements. A basic crowbar circuit is shown in Figure 5-19 . In the event of a tube fault, a signal derived from the fault current sensing element activates the trigger unit, firing a crowbar element which places a very low impedance (arc) across the power supply leads. Simultaneously, a "turn-off" signal is fed back to shut off the power supply. The resulting energy dissipated in the tube arc is only a few joules; most of the energy stored in the power supply is dissipated in the crowbar arc. Two resistors are normally used in the circuit as shown in Figure 5-19. Resistor R-1 limits peak fault current to a safe value which precludes damage to the energy storage capacitor and the crowbar device. Resistor R-2 insures that a high percentage of the fault current will flow through the crowbar device rather than through the tube. The trigger unit normally contains a repetitive firing feature, insuring that a significant charge cannot reappear in the energy storage capacitor in the event the crowbar device extinguishes before the power supply is turned off.

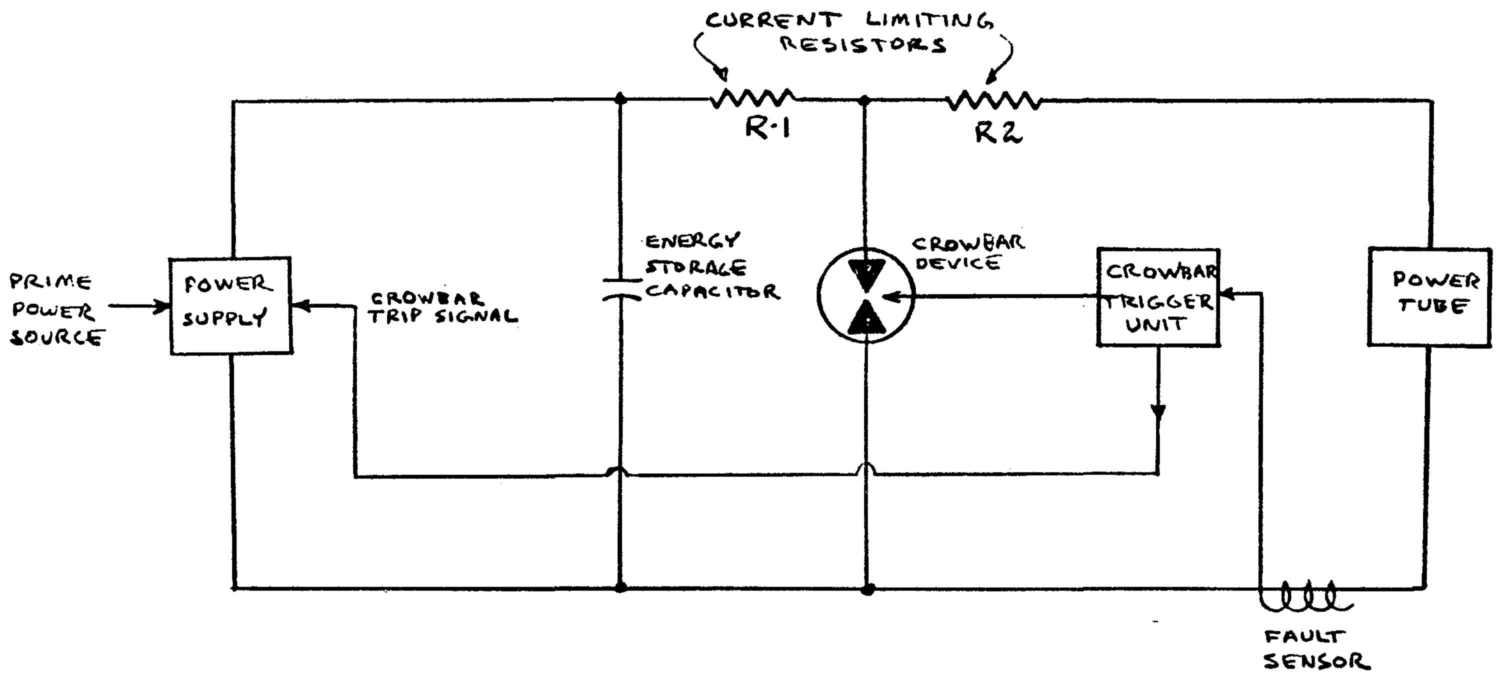


Figure 5-19. Basic Crowbar Circuit

5.5.2 Requirements

The basic requirements for the monitor and protective subsystem are outlined in Appendix A-5. Specific circuits required are "electronic crowbars" for the plate (anode) high voltage supplies of all of the high power amplifiers. As an integral part of the crowbar protection, a fault sensing and control logic circuit is required to sense faults potentially harmful to the final amplifier components, fire the crowbar element, and turn off the power supply to limit fault energy. In addition, a VSWR trip is required for RF fault protection. The output of the task was to provide a breadboard crowbar design, along with associated monitoring, control, and protective circuitry to complete the fault protection system.

The crowbar protective system will be capable of maintaining the fault arc energy below a level of 5 joules in the case of an arc in one of the Doherty final amplifier tubes. This value has been established as a reasonable limit of the energy the planar triode in the amplifier may safely endure without permanent damage or disability. The power conditioner has an L-C section output filter which stores the energy required in providing sync peak power for the transmitter. The present engineering estimate of this filter is a 1.6 Hy choke and a 45 μf capacitor. The crowbar element with 2.5 kV must be capable of handling an energy of

$$\begin{aligned} W_c &= 1/2\ CV^2 \\ &= 140 \text{ joules} \end{aligned}$$

Spark gaps are available for operation over a side range of energy requirements, and this level presents no problem as far as rating is concerned.

The power supply with crowbar-protected load configuration is shown in Figure 5-20 .

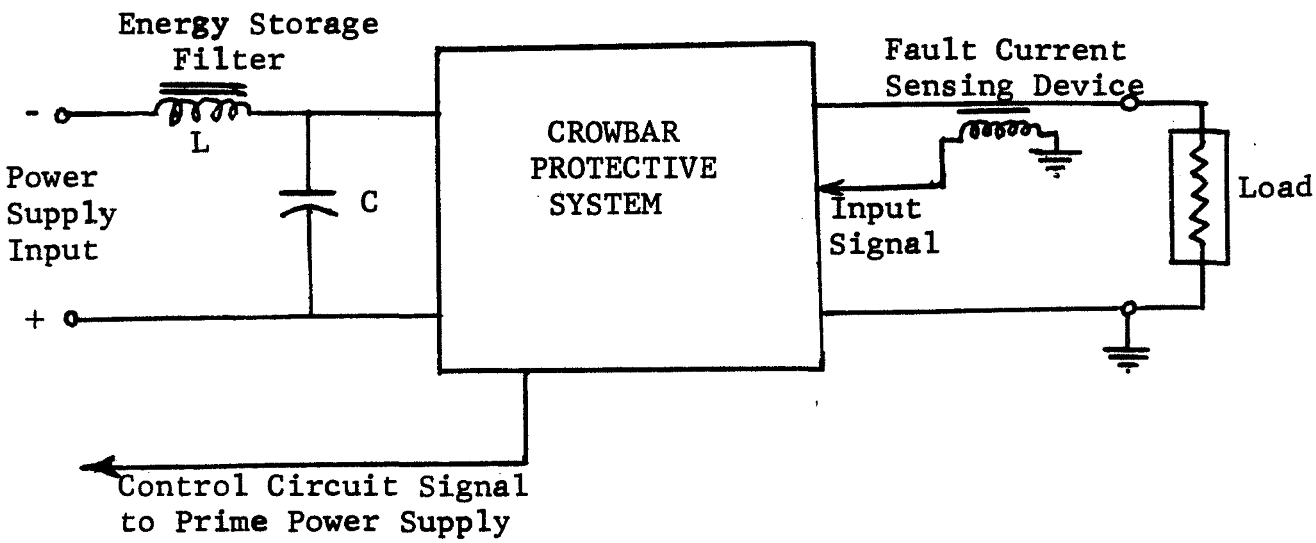


Figure 5-20 . POWER SUPPLY - CROWBAR - LOAD DIAGRAM

This circuit was analyzed to determine the division of energy when an arc appears in the load (protected tube) and the crowbar gap is triggered. The crowbar gap can be triggered from the current transient to the load, and generally the circuit can divert the current in well under 2 microseconds. In the case computed, the fault energy was about 0.1 joule, well below the level that would be injurious to the transmitter tubes.

In order to realize this performance, the internal impedance of the storage capacitor must be less than about 250 ohms to insure a minimum keep-alive current. In addition, the storage capacitor must be of a quality that will permit a large transient current. Repetitive triggering of the arc is built into the circuitry to insure that the capacitor will not recharge from energy stored internally in the power conditioner and in the filter choke after initial arc extinction.

5.5.3 Crowbar for Doherty Amplifier

5.5.3-1 Crowbar Devices

A survey of types of crowbar devices was made, and their relative merits are compared in Table 5-6 . The vacuum spark gap was chosen initially for the high voltage power supply, but inconsistent operation eventually led to using a gas filled type. The filter in the power supply contains a large energy storage capacitor, and the ruggedness of the spark gap must be sufficient for a long gap life. The relatively simpler Krytron* is a good choice for use in smaller power supplies.

Available vacuum gap sizes fall on either side of the required value. The larger of the two sizes closest to the design value of 140 joules was chosen initially for reasonable gap life; this was an EG&G GP-12BV. The GP-12BV, a vacuum version of the GP-12A, has an operating range of 1 kV to 50 kV in air with a peak current capability of 100,000 amperes. It requires a trigger potential of 20-30 kV and will fire after a delay of only .05 μsec. The total dissipation rating of the gap is 2500 joules.

As noted previously, this oversized gap did not fire consistently when operated well under its rating, so the GP-20AV was substituted. This was rated at 1.0 to 11.0 kV with a peak current capability of 15,000 amperes, and a dissipation rating of 200 joules. This unit also acted erratically, and subsequently the gas filled GP-31A tube, also a 200 joule tube, was used with good reliability in firing. It has the disadvantage of a smaller dynamic voltage range, about 2.0 to 6.0 kV, but operates with up to 15,000 amperes of current. However, it showed good stability and was judged to be superior for the Doherty amplifier where consistent firing is essential. The test program approach was subsequently modified to take into account

* Trademark of EG&G

Table 5-6. Crowbar Device Comparison

TYPE	TYPICAL EXAMPLE	ADVANTAGES	DISADVANTAGES
1) Spark Gap	Gas Filled Type	Can handle very large currents for longer times, simple rugged construction -- requires no keep alive current.	Possibility of leakage of ionizing gas.
	Vacuum Gap	Same as gas filled gap + wider operating voltage range than gas gap.	Erosion and sputtering cause failures, require currents on order of 1 amp to maintain conduction, therefore repetitive firing may be necessary.
2) Cold Cathode Switch Tubes	Krytron (EG&G)	Very small size and fairly large current handling capabilities. Very little delay (ns) and fast ionizing time (μsec). Low voltage trigger, will conduct at current levels down to milliamperes.	Glass envelope construction, ionizing gas filled keep alive current. Generally low energy capability $\leq$50 joule life limited by cathode deterioration. Keep alive current contributes to gas clean up.
3) Hot Cathode Switch Tubes	Hydrogen Thyratron	Fast firing and ionizing (on order of 1 μsec)	Requires high amounts of heater power (up to 50 W) for larger versions, some are glass envelope (vacuum tube) types.
4) Solid State	Silicon Controlled Rectifier	Requires no heater/filament power, higher reliability by virtue of solid state properties (no vacuum seal or glass envelope)	Highest voltage rating presently available about 1700-2000 volts DC, slower turn on time depending on gate characteristics ($\geq$2 μsec). Higher voltage ranges require series connections of these devices with attendant reduction of reliability.

the minimum voltage capability of the GP-31A.

The generalized circuit of Figure 5-19 was used for all crowbar circuits. After the device was selected, the auxiliary circuitry was arranged to adapt to its requirements and capabilities.

5.5.3-2 Trigger, Logic, and Control Circuits

The crowbar trigger, logic and control circuitry, shown in Figure 5-21 consists of a threshold circuit employing Q_1 to trigger the monostable multivibrator composed of Q_2 and Q_3. The pulse length of the latter is varied by changing the value of the 5K pot in the collector of Q_2. Pulse lengths of up to 80 msec can be achieved. The output pulse is fed simultaneously to trigger an astable multivibrator and to a relay driver circuit (Q_9, Q_{10}, and Q_{11}). The relay provides a disconnect signal for the prime power source while the free running multivibrator (Q_4, Q_5, and Q_6) and associated amplifier (Q_7 and Q_8) provide output pulses of 10 volt amplitude for the length of time the one-shot multivibrator is on. The pulse repetition rate is approximately 30 pps; up to 4 pulses are available depending on the one-shot multivibrator adjustment. The output pulses are fed to the TM-11 commercial trigger module (see Figure 3-26) which develops 30 kV pulses to fire the spark gap.

5.5.3-3 Trigger Module

An EG&G TM-11 Trigger Module was used to provide the 30 kV pulses necessary to fire the GP-12BV vacuum spark gap, and later the GP-31A gas spark gap. The module consists of a DC power supply, pulse transformer and KN2 Krytron discharge tube which discharges a variable amplitude pulse into the pulse transformer to provide a firing pulse variable in amplitude from 15 to 30 kV. The module can be triggered from a front panel switch, a remote switch, or by 10 V pulses from a pulse generator.

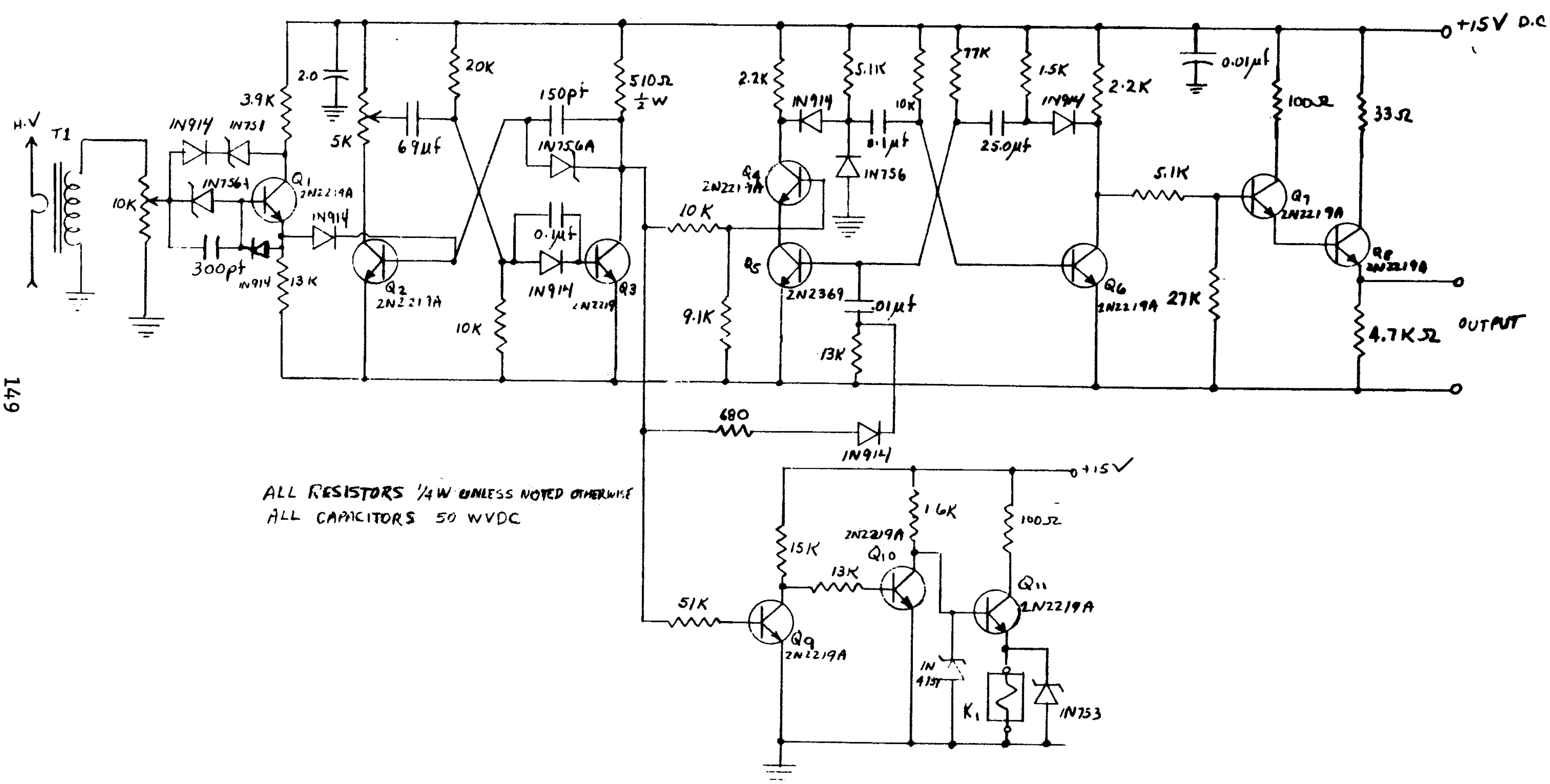
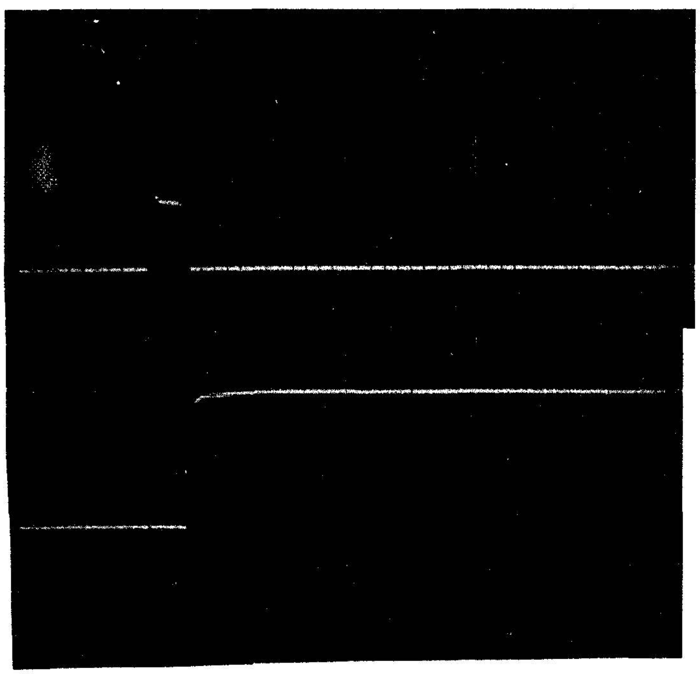


FIGURE 5-21. TRIGGER, LOGIC, AND CONTROL CIRCUITRY

5.5.3-4 Operation

The crowbar unit is placed in operation by providing interconnections to power supply control circuits such that the closure of a normally open pair of relay contacts will remove high voltage. High voltage connections are made through special BNC style connectors on the chassis rear surface. An indication of the speed of action and degree of protection provided by the crowbar was the aluminum foil, test, shown in Figure 3-28. The total delay time between the fault current pulse and the first output 30 kV trigger pulse of the TM-11 has been measured to be 0.6 μsec, and is shown in Figure 5-22. The delay from the input pulse until the relay closed was about 3 microseconds.



VERT = 5 Volt/cm

HORIZ = 1 μsec/cm

Top trace is input pulse to trigger circuitry.

Bottom trace is front of firing pulse to TM 11 module.

Total delay shown = 0.6 μsec

FIGURE 5-22. MEASURED TRIGGER PULSE DELAY

5.5.4 VSWR Trip Circuit

Protection of the transmitter output tubes against dangerously high VSWR's is provided by a circuit which samples the reflected rf signal from a directional coupler, rectifies it, and compares it to a present level. If the rectified signal exceeds

this threshold a relay is actuated which in turn removes power from the transmitter and thus protects the output tubes. The schematic of this circuit is shown in Figure 5-23.

5.5.5 Driver Crowbar Design

A schematic diagram of the crowbar used in the 1 kV power supply for the visual driver amplifier stage is in Figure 5-24. It employs a KN2 Krytron tube as the crowbar element and a TR149 trigger transformer to fire the tube. The crowbar fires with very little delay, successfully passing the 0.5 mil aluminum foil puncture test.

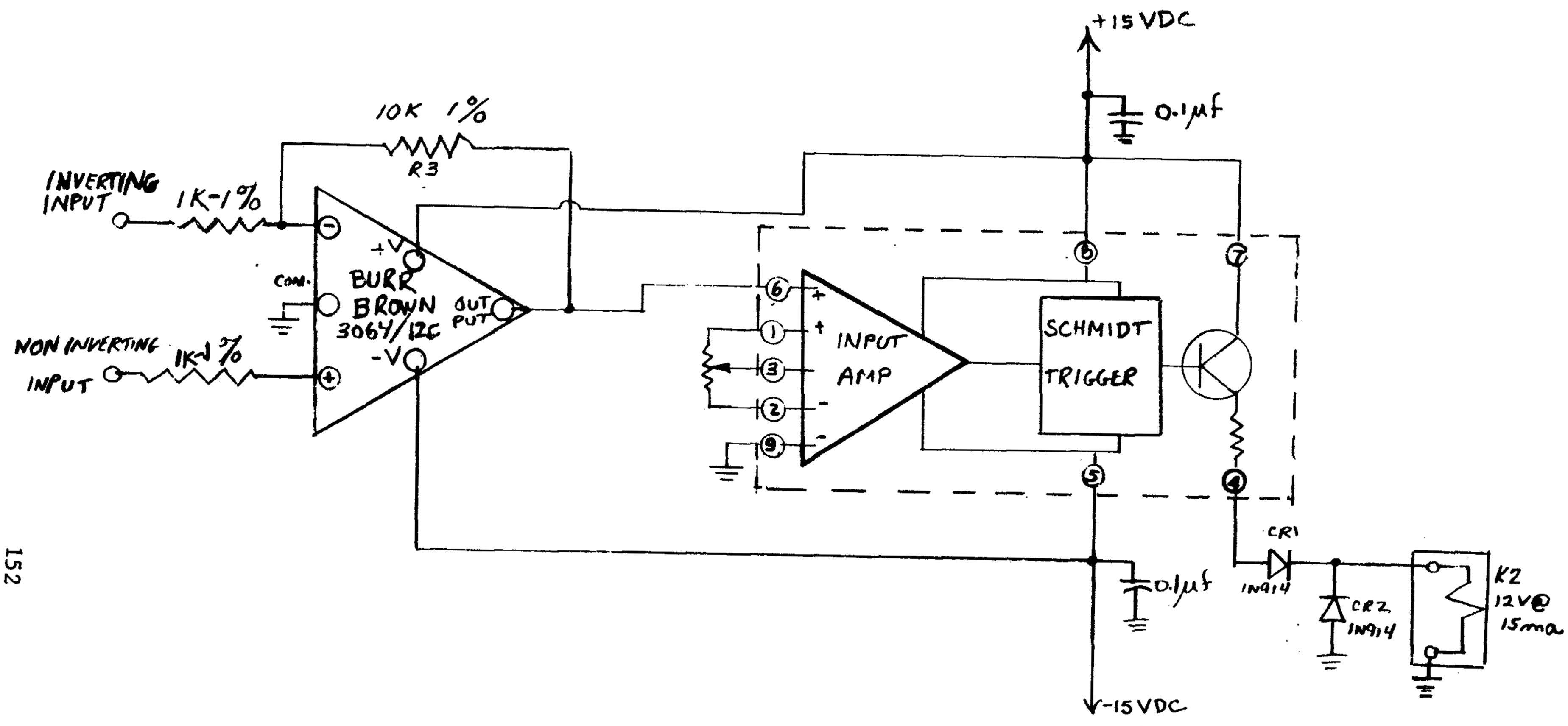


FIGURE 5-23. VSWR TRIP CIRCUIT

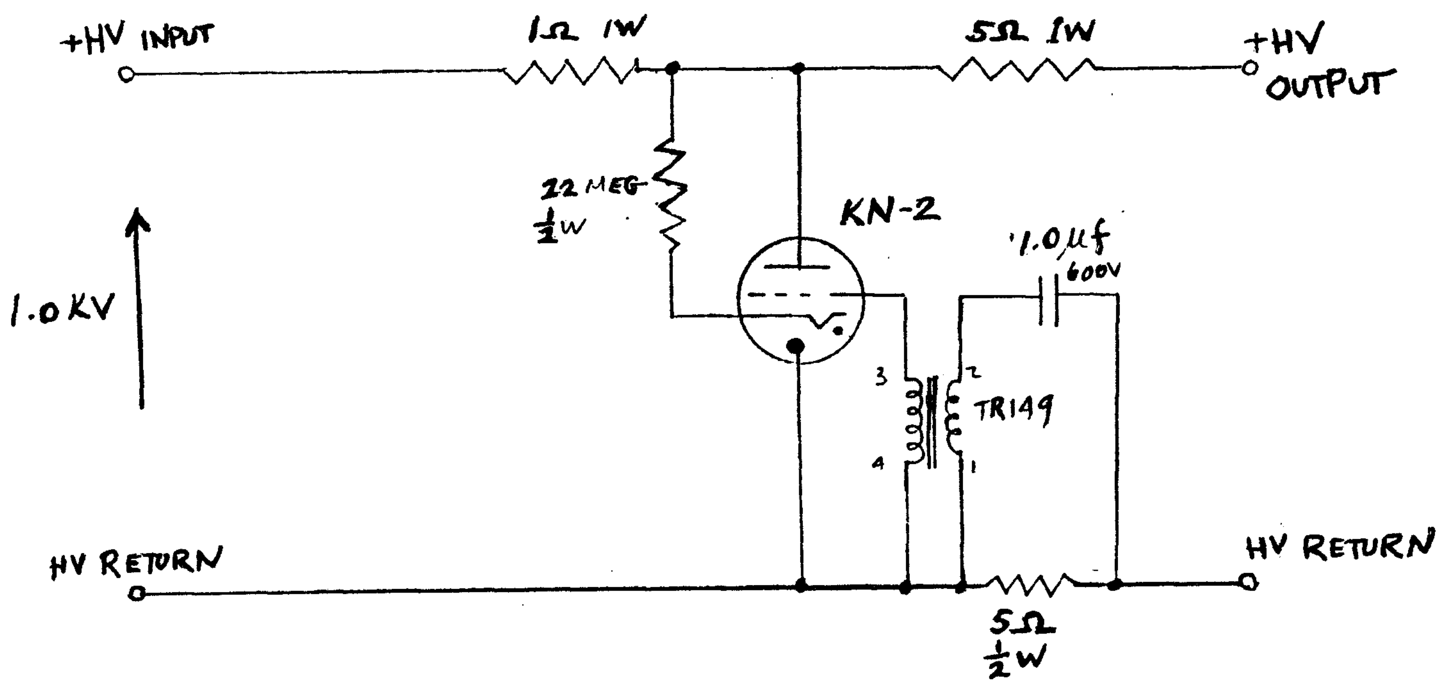


FIGURE 5-24. DRIVER CIRCUIT CROWBAR

5.6 CONTROLLED CARRIER CIRCUIT DESIGN - TASK 6

5.6.1 Systems Aspects

A company-funded development[7] evaluated a proposed technique of amplitude modulated television transmitter carrier control (a form of Automatic Level Control) for more efficiently utilizing the spacecraft prime power source capability. The need for this feature arises from the fact that the average energy content of a U.S. Standard composite TV signal varies over a 4.8 dB range for a fixed sync peak power level, and the dc power required for a high efficiency amplifier follows the average RF power variation. An earlier system study[12] indicated that optimum dc power systems for space broadcasting will not have a battery storage capability for the high power transmitter, so the prime power system would be sized for maximum average power requirement, which occurs for a black picture in the US and most other TV systems. The subsequent study indicated that the sync peak output level of a simulated high efficiency transmitter could be regulated so as to maintain prime power requirement constant at the average gray picture level about (32% duty) with no discernible effect to the average viewer as long as the level controlling circuit has a time constant response the same as or somewhat longer than the receiver AGC circuit. It is then possible to size the power source for an average gray level power demand rather than for a maximum black level video. The savings in prime power would typically be about 30% for the Doherty amplifier circuit.

A basic carrier control circuit modeled in the above mentioned program is shown in Figure 5-25 . Basically, this circuit varies the RF drive signal amplitude such that the average power in the transmitted signal is constant. Thus, the sync peak power would be high for a white picture which requires relatively little picture power, and would be low for an all black picture which requires a high picture power. Variation of sync peak might be about 2.4 dB. This is not the best circuit operation, however, since the peak power rating of the transmitter would be keyed to the white picture level, and it would be operating well below its rating most of

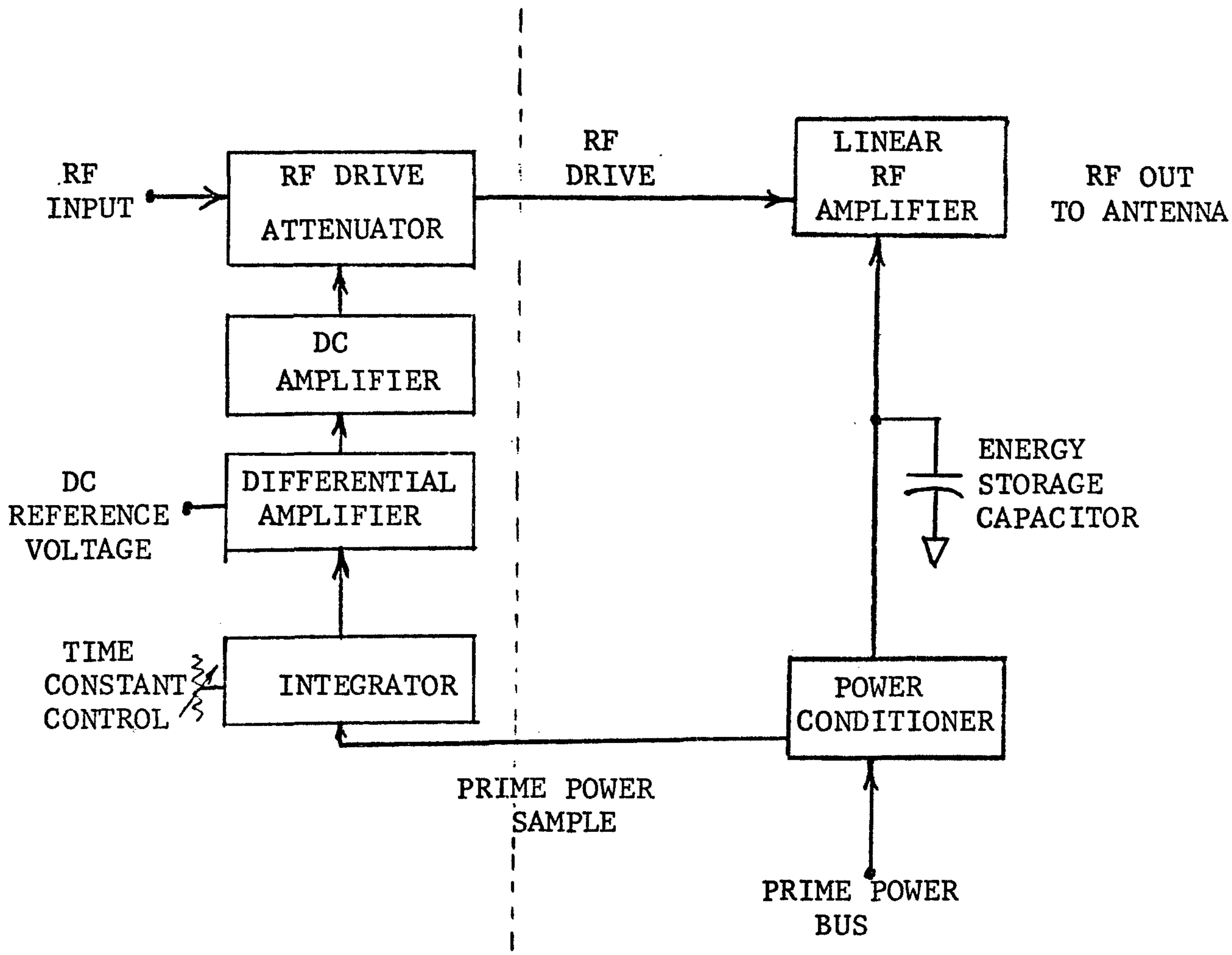


FIGURE 5-25. BLOCK DIAGRAM OF CONTROLLED CARRIER CIRCUIT

the time, resulting in low efficiency and partially defeating the purpose of the controlled circuit. The alternate proposed circuit, which is developed in this program, would use an average gray level threshold such that the amplifier has a maximum output for that picture level. Then, the sync peak would remain constant for all white pictures below threshold gray (with a corresponding reduction of required dc power but with good efficiency), but would decrease for the larger dc power requirements of dark pictures. In the latter situation, the carrier control circuit would keep the total average power constant. Thus, the amplifier will be operating more closely to its best efficiency point with dark pictures than would the continuous functioning circuit without the threshold feature.

Viewing tests have only been performed for the continuously varying drive signal circuit. Viewing tests should be performed also with the threshold circuit, although there is no reason to suspect a significant difference from normal performance.

5.6.2 Requirements

The basic requirements for the controlled carrier circuit are in Appendix A-6; supplemental requirements include the following:

<u>RF Gain Control Element</u>

Insertion Loss	1.0 dB or less (minimum attenuation) to 6.0 dB (maximum attenuation)
Signal Transfer Amplitude Linearity	$\pm$ 0.25 dB over the video signal range of + 0 to -20 dB (referred to sync peak) at any point of the 6.0 dB variable loss range
Phase Linearity	$\pm$ 3° over the video signal range of + 0 to -20 dB (referred to sync peak) at any point of the 6.0 dB variable loss range
RF Connectors	Type N

<u>Power Demand Sampler</u>

Sampler Element	Sampling resistor in B- circuit; followed by adjustable threshold circuit.
Sampler Output	A dc voltage proportional to instantaneous power demand of rf amplifier.

Sampler Output Voltages	Typically 1.0 to 5.0 volts average level at the control point
<u>Integrator and Amplifiers</u>	
Input voltage	From power demand sampler
Output	As required to drive the RF Attenuator Element over the specified range.
Power Demand Regulation	$\pm$ 2% maximum deviation between the "average gray" clamp level and the "all black" video signal
Average Gray Clamp	A selected average video level in the composite TV signal, gain must be constant for signals below the clamp level.
Control time constant	Variable, 0.6 to 20 milliseconds
Drift in Gain	$\pm$ 0.1 dB over a 4 hour period for a normal air conditioned room environment ($\pm$ 3°C)
Test Points	Monitoring points for all significant currents and voltage
<u>RF Control Element</u>	Diodes packaged in an rf shielded enclosure
<u>Personnel Safety</u>	
High Voltage	All terminals more than 24 vrms above ground will be adequately insulated or shielded
RF Radiation	Maintained below 10 milliwatts per square centimeter at all points accessible to personnel
Hot Spot temperature	The outside surfaces of circuitry operating above 100°C will be adequately shielded to prevent personnel contact.

5.6.3 Circuit Design

The circuit of Figure 5-26 is divided into four components : the RF Drive Attenuator; the Attenuator Control circuit including the DC Amplifier, Differential amplifiers, and Integrator; the Power Sampling Circuit; and the HV Power Supply terminals. The Power Sampling is achieved by monitoring the final amplifier plate current as determined by the voltage developed across a 2 ohm resistor (R1) located in the high voltage power supply return. Normally, the average dc plate current will be 1.5 amperes; therefore, the nominal monitoring voltage will be 3 volts. Voltage tran-

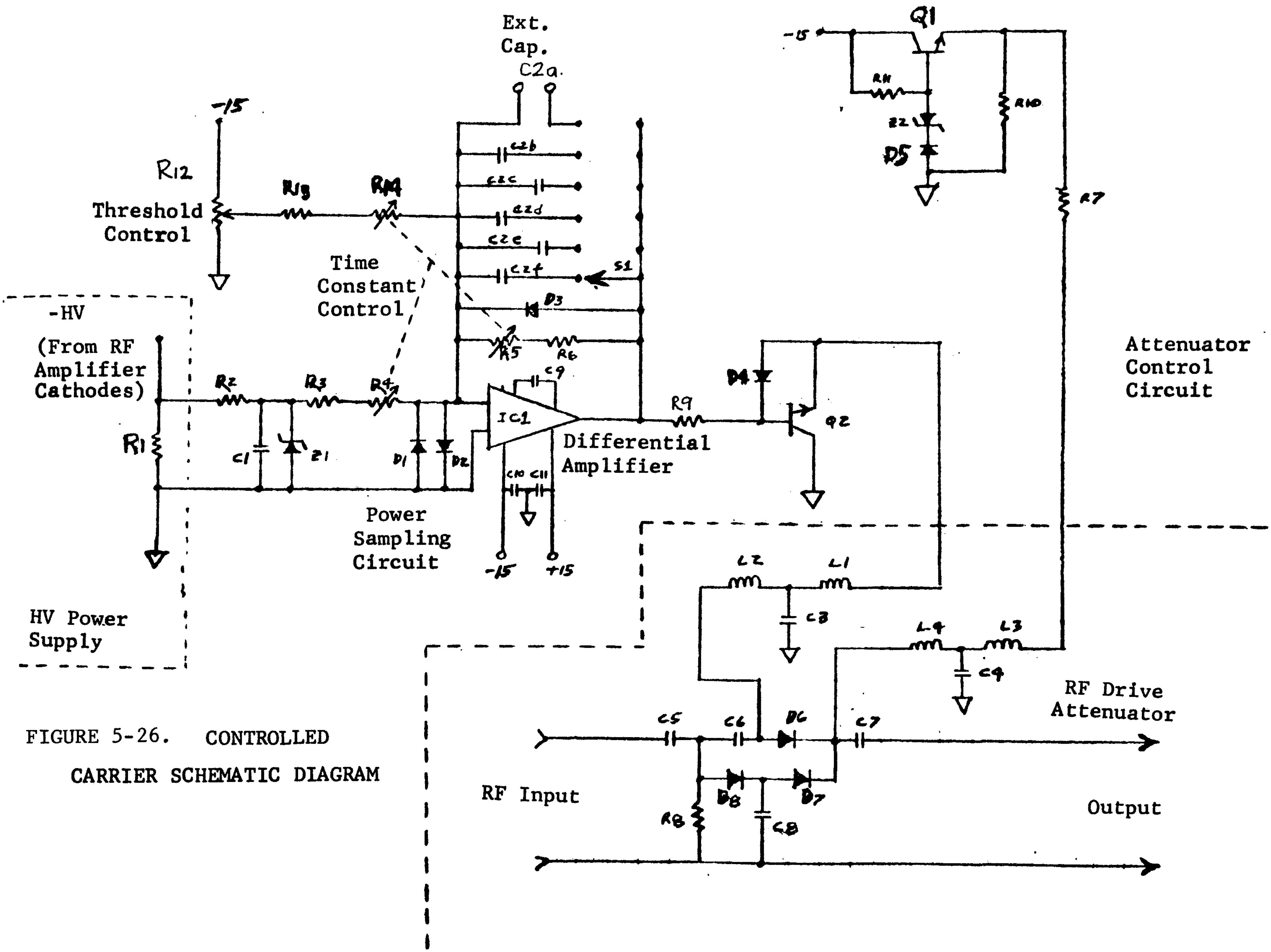


FIGURE 5-26. CONTROLLED CARRIER SCHEMATIC DIAGRAM

sients that appear across this resistor will be an indication of the current or power demand changes at the final amplifier stages.

The Attenuator Control Circuitry amplifies and actively filters the sampled voltage of the power sampling circuit and drives the RF attenuator. The Sampling Circuit input network composed of C1, Z1, D1, D2, R2, and R3 serves as a low pass filter with a break point of 310 Hz (3.23 msec). The network also protects the front end of the IC amplifier (IC1 in Figure 5-26) from any transients that may appear. The control threshold level is set by R12, R13, and R14. The time constant of the feedback network is varied by R5 and by switching C2; six ranges cover a 0.24 to 22 msec. span. Provisions are included for adding an external capacitor to the filter network. The three potentiometers are ganged to keep the gain and threshold constant while varying the time constant.

Diode D3 clamps the output of the amplifier at + 0.7 volts maximum, thus limiting its dissipation during periods of minimum attenuation. Transistor Q2 serves as an emitter follower to handle the required attenuator drive current.

A compensation regulated -10 volts is supplied by the series regulator Q1, Z2, D5, R10, and R11. Diode D5 compensates for the V_{be} due to temperature. The RF Drive Attenuator uses three PIN diodes in a pi configuration, biased in the proper ratios to keep the input/output impedance constant at 50 ohms over the usable range of 0.5 dB to 6 dB. In the full-on state, the emitter of Q2 is at approximately zero volts, and a 50 ma bias current flows through the series element D6 while the shunt elements D7 and D8 are reverse biased. As high attenuation is demanded, the emitter of Q2 goes more negative, reducing the current through the series diode D6 and increasing the current through the shunt diodes, D7 and D8. R7 and R8 are chosen to provide the proper bias ratio for constant impedance.

The two filters L1, L2, C3, and L3, L4, C4 are low pass isolation filters with break points of approximately 10 MHz. Capacitors C5, C6, C7, and C8 are dc blocking capacitors with an impedance less than 0.1 ohm at 800 MHz.

The circuit provides the required performance, and was fabricated for tests on the complete breadboard transmitter.

5.6.4 Implementation

The general analysis of the circuit showed no difficulties in the circuit approach. Voltages, currents, impedances, and required components are all within normal ranges. The parts required for the circuit of Figure 5-26 are:

R1	2 ohms, 25 watts, non-inductive		
R2	270	C1	.5 µf
R3	270	C2a	external
R4*	500 per section	C2b	.01 µf
R5*	50K per section	C2c	.022 µf
R6	51K	C2d	.047 µf
R7	180 1 watt	C2e	.1 µf
R8	62K	C2f	.22 µf
R9	510	C3	.002 µf
R10	2K	C4	.002 µf
R11	330	C5	.002 µf
R12	2K pot.	C6	.002 µf
R13	100K	C7	.002 µf
R14*	100K pot.	C8	.002 µf
		C9	30 pf
* Ganged - all linear taper		C10	.01 µf
		C11	.01 µf

Q1 and Q2	2N1711
D1 to D5	1N914
D6 to D8	UM6006 PIN
IC1	LM101A
L1 to L4	10 μh

5.6.5 Tests

The circuit was completed but system tests were deferred. It was designed to operate with the high power transmitter, and would not provide significant performance data at the low power levels used in the transmitter tests of Section 5.8. The circuit is available for inclusion in tests when the higher power tubes are obtained.

5.7 HIGH POWER RF COMPONENT ENVIRONMENTAL TESTS - TASK 7

5.7.1 Approach to Tests

This task was to implement testing of several UHF RF components in a vacuum environment. In descending order of preference, the several representative test items included in the test plan were:

1. Uniform 3-1/8 inch coaxial line	3-1/8 inch, 50 ohm
2. Uniform Waveguide line	Half-height WR975
3. Coaxial step, 3-1/8 inch, Length = 14 inches	a) 1 mm gap b) 2 mm gap c) Contaminated Surface (Oil film) d) Plasma Torch Spray Materials
4. Waveguide Step, half-height WR975 Length = 20 inches	a) 1 mm gap b) 2 mm gap c) Contaminated Surface (Oil film) d) Plasma Torch-Spray Materials
5. Color Subcarrier Image Filter-Waveguide	WR975 Half-height, 10" x 14"
6. 3 dB Hybrid, Dual WR975 waveguide	Length - 24 inches

Of particular interest but not vital to the immediate program are:

7. Waffle Iron Section	10" x 20"
8. Waveguide Slot	10" x 12"

The two stepped sections (items 3 and 4, shown in Figures 3-33(b) and (d)) represent segments of reactive harmonic filters, which may be used in a high power system for coaxial and waveguide transmission. They also represent lower impedance coaxial, rectangular waveguide, and ridged waveguide lines which are possible approaches to multipactor prevention. (See Appendix B for low impedance line analysis.) The two items with step sections are matched by quarter-wavelength impedance transformers. Each step section component was fabricated with a gap separation of one millimeter (.039") to minimize probability of multipacting; additional units of wider gap spacings might be considered to extend these types of measurements in future testing.

6

The waveguide stepped section is shown in Figure 5-27.

An assortment of difficulties was encountered as testing efforts progressed. The results were limited ultimately by the lack of an adequate high power rf source; the original 2.5 kW source became faulty and the 500 watt aural stage was used as a substitute. However, it was not capable of achieving the higher power levels (on a pulse basis) with the available Y1498 tubes. The overall facility and its problems are outlined below. The original rf power source, which used an early research version of the Y2042 tube, operated during tests on the standard coaxial line (item 1) which showed no breakdown tendencies (as expected). With the failures of the research tubes, effort was expended in using the aural stage. Tests on the other components were subsequently deferred until a better set of tubes was obtained, and at least a 2.5 kW rf signal could be realized.

5.7.2 Test Facility Description

A facility was assembled for the testing of rf breakdown performance of the several typical UHF transmission components in a high vacuum, simulating conditions expected in spacecraft applications of these components. The basic test circuit is shown in the simplified block diagram of Figure 5-28. It consists of an rf source, nominally rated at 2.5 kW CW at 800 MHz which transmits rf energy through a vacuum window into the component under test which is mounted in the vacuum chamber. The rf exiting from the component passes through a second chamber window and into a high power load which is mounted outside the chamber. The component being tested is vented into the vacuum chamber so its internal pressure is expected to approximate that of the chamber. Breakdown performance of the component is monitored by observing forward and reverse rf power flow, chamber and component internal vacuum levels, component temperatures, and multipactor breakdown sensor output.

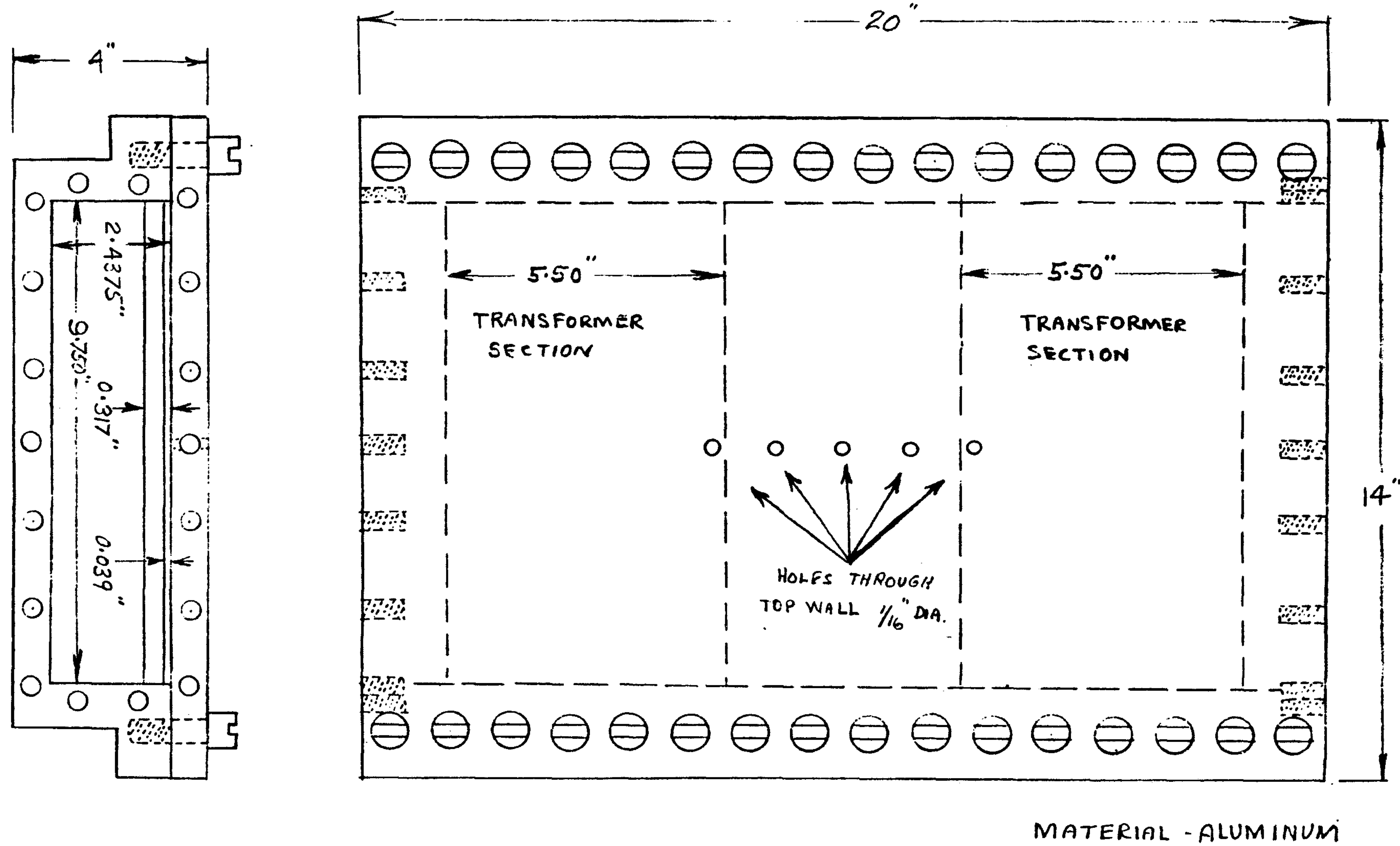


FIGURE 5-27. STEPPED WAVEGUIDE SECTION

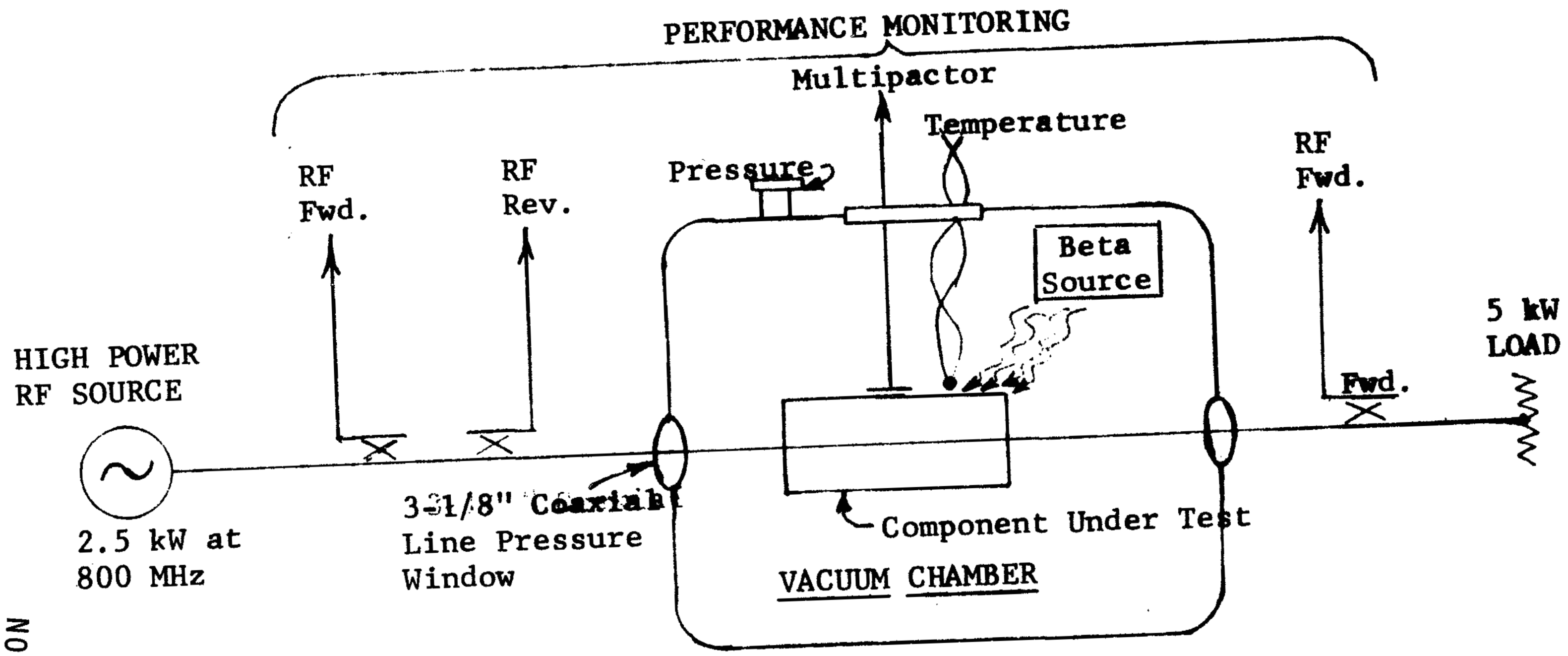


FIGURE 5-28. BASIC RF BREAKDOWN TESTSCIRCUIT

A detailed circuit in Figure 3-32 , shows the final rf source test configuration. An earlier version used a different high power rf source (as noted previously and explained in a later paragraph , but was functionally the same. Problems with the power amplifier tubes used in both rf sources precluded operation at the intended 2500 watt level. The overall setup is seen in the Photograph of Figure 5-29. RF power is fed from the power amplifier through the circulator and 7/8" Heliax* coaxial cable, through the coaxial vacuum window, through the test piece, and out a second vacuum window. A forward power monitor directional coupler and a coaxial load mounted above the chamber terminate the train of components.

The interior of the chamber with the 3-1/8 inch coaxial line test section in place is shown in Figure 5-30. Indicated in the photograph are the three multipactor sensor electrodes which are distributed along the length of the coaxial step section where multipactor breakdown was likely to occur at a power level below 2.5 kW. The 200 MHz multipactor electrode and matching tank circuit are also shown. The 200 MHz circuit was provided to inject free electrons into the test section to insure a high probability of multipactor occurrance within a short time if the proper conditions are present within the section under test. (A beta particle source was also incorporated for the same purpose as a backup ionizing source.) Components of this 200 MHz assembly were also used in a multipactor suppression test as discussed later in section 5.7.4. Also visible are a number of thermocouples for monitoring temperature distribution along the rf assembly. Note the large number of holes drilled into the coaxial outer conductor to insure adequate venting of the coaxial line interior to the chamber. An ionization gauge (not visible) was also coupled to the interior of the coaxial line for internal pressure monitoring. All areas of the rf circuit which might provide "gas traps" were drilled to insure venting to chamber pressure. Considerable care

* T.M. Andrew Corporation, Chicago, Illinois

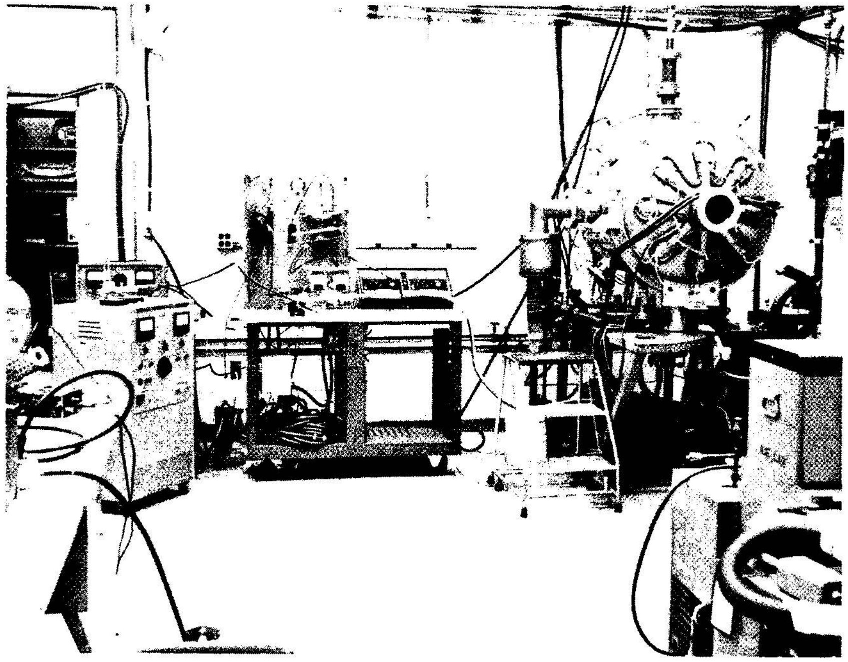


FIGURE 5-29. RF BREAKDOWN TEST FACILITY

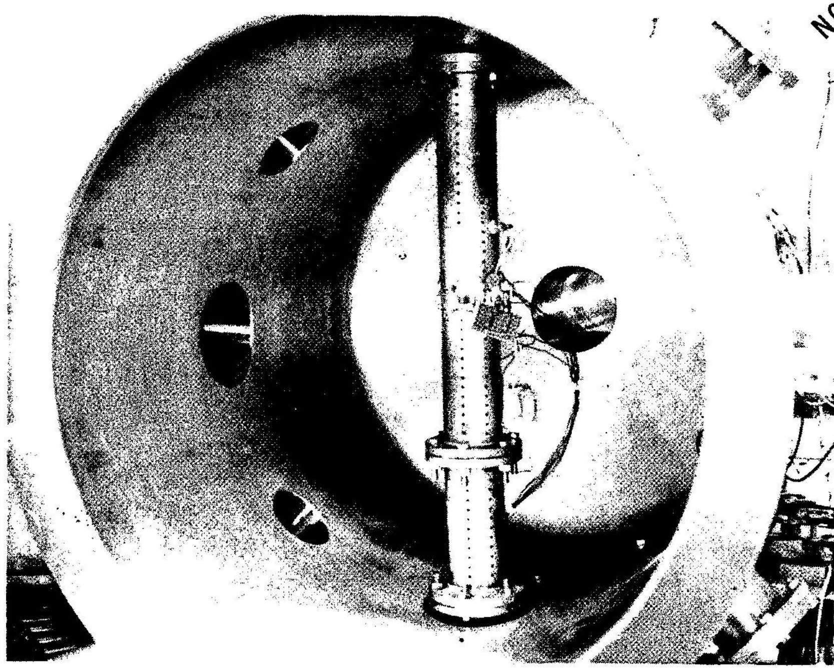


FIGURE 5-30. TEST SECTION IN VACUUM CHAMBER

was exercised in the alignment of test components and chamber, and in leak control measures so that pressure of 10^{-6} Torr could be obtained.

5.7.3 Test Components

In addition to testing the 3-1/8 inch coaxial line and the half-height WR975 waveguide, several components were designed and fabricated for breakdown testing. These components, illustrated in Figure 3-33 are the coaxial and waveguide step, the 3 dB hybrid-sidewall coupler, and the waveguide cavity filter. The two step sections were deliberately designed to break down below 2.5 kW so that the effects of multipactor breakdown (i.e, in sensor development) and of multipactor suppression schemes could be evaluated. These step sections are also representative of several reflective filter types.

The 3 dB hybrid is a component used in many applications such as in some types of directional filters (used in frequency multiplexing) and in power combiners. The cavity filter was intended to be operated at Q levels which would give voltages equivalent to that experienced in the aural notch filter used in the TV transmitter breadboard circuit. Photographs of these components are shown in Figure 5-31 . (The cavity filter uses the waveguide shell which is also used for the step breakdown section and thus is not shown separately.)

5.7.4 Test Results

Initial operation of the system shown in Figure 5-28 was achieved with the coaxial line installed, using the 2.5 kW test transmitter developed on an earlier General Electric funded program. This unit was designed to test and evaluate the L64S and L67E triodes which were forerunners of the Y1498 and Y2042. Two L67E tubes were furnished for use in the power amplifier stage.

No evidence of breakdown or excessive outgassing was noted at test power levels up to 1500 watts CW under hard vacuum conditions. At this point, the first L67E

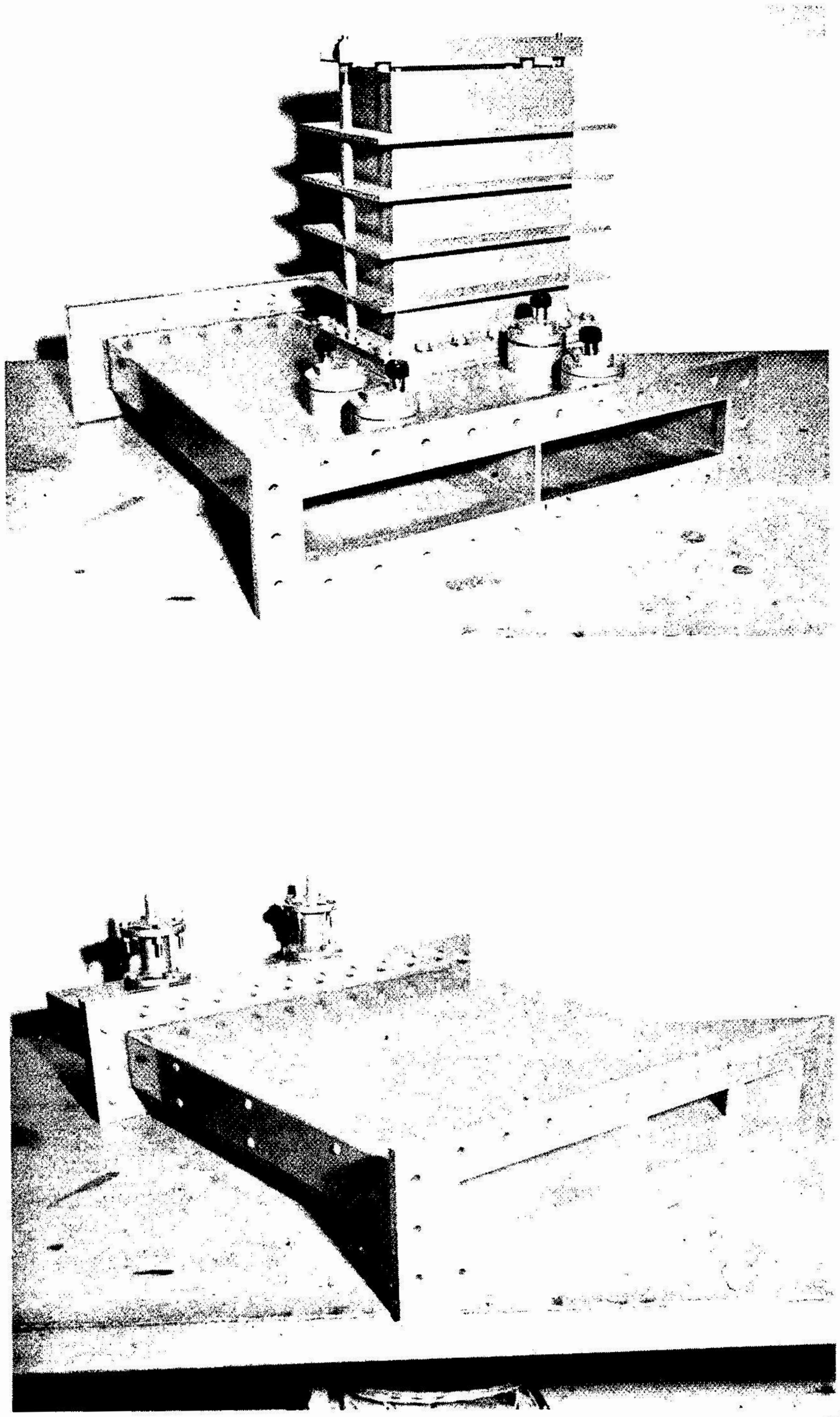


FIGURE 5-31. AURAL NOTCH FILTER AND 3-dB HYBRID

failed with a grid cathode short. The other L67E failed in the same manner at several hundred watts output. No additional L67E tubes were available and the equivalent Y2042 was not yet available from the production facility at Owensboro, Kentucky. The aural amplifier stage, after initial testing, was put into service for further breakdown testing. The Y1498 oxide cathode tube, however, would not operate much above 500 watts CW without overheating the grid, and the gain was too low to achieve more than 700 watts peak output because of driver power limitations. Further high power testing was then suspended; the higher power Doherty amplifier, when available from the breadboard transmitter, would provide the required operation.

A significant test was performed on a multipactor suppression scheme which was conceived during the course of the program. In a number of cases, the use of screen or mesh electrodes has been used in parallel-plate multipactor test setups to eliminate normal multipactor modes so that the presence and characteristics of "one-sided multipaction" could be evaluated. Since single-sided multipacting occurs only when a certain minimun dc bias (electric or magnetic field) exists between plates (Figure 5-32), the mesh surface on one plate effectively eliminated all multipactor tendencies as long as zero bias exists between the electrode plates. The question, therefore, is whether this suppression technique is practical and does it avoid significant side effects, such as greatly increased RF dissipation and complication of RF leakage through the perforations. If too much perforation is employed, RF losses will be high; too little would provide marginal performance.

A preliminary examination of the RF loss indicates it should not be a serious factor. To illustrate, a waveguide section with broadwalls perforated is shown in Figure 5-33 . Assume that the electric field has accelerated a number of free electrons from the left wall vicinity towrad the right wall; some will pass through the holes in that surface, as in the case of the Faraday cup sensor for multipactor

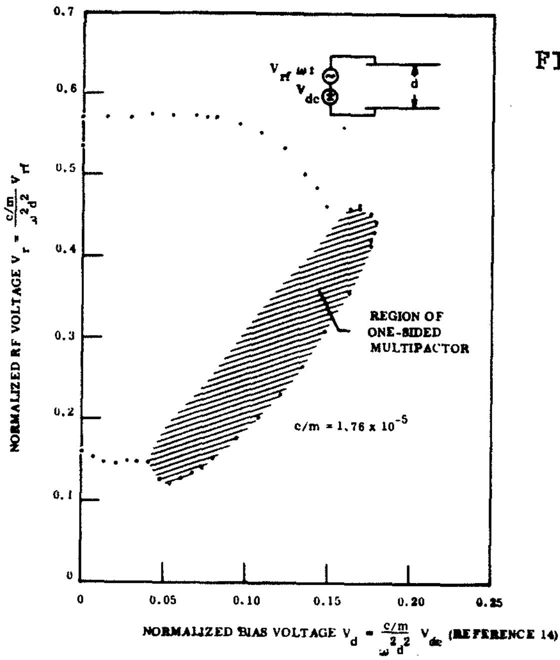


FIGURE 5-32. REGION OF MULTIPACTOR DISCHARGE WITH DC BIAS

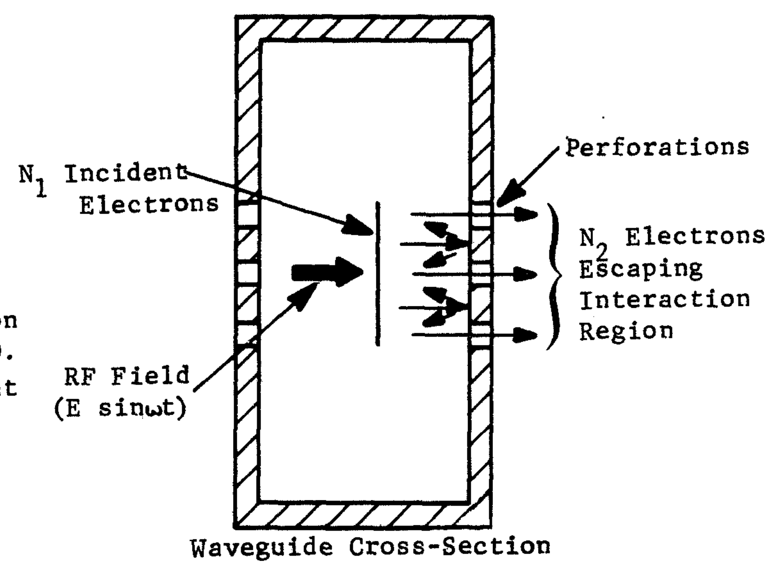


FIGURE 5-33. MULTIPACTOR SUPPRESSION WITH PERFORATED WAVEGUIDE

(Note: Multipactor suppression achieved when $(N_1-N_2)/N_1$ 1.0. Secondary emission coefficient of surface effectively less than 1.0.)

breakdown. If the maximum secondary electron emission coefficient of a metal such as copper is only about 1.3, then 25% of the incident electrons mush be removed, implying that 25% of the top and bottom surfaces must be removed. The sidewalls and other areas not subject to multipacting could remain solid since they do not contribute to the breakdown process. Since perforation is not required in areas of components where the $V/(fd)^2$ is not appropriate to support multipactor breakdown, the removed area may turn out to be between 1% and 15% of the total waveguide system area. This would increase RF loss to less than 0.1 dB for the system.

The effectiveness of the perforated plate scheme was evaluated at 200 MHz using the test setup of Figure 5-34. With both 1-inch by 3-inch plates of solid aluminum, multipactor discharge was noted at 0.70 watts input with 1/2-inch gap between plates. With one plate perforated with twenty-five 1/4-inch diameter holes (41% area removed) the power required to start multipaction increased to 7.0 watts.

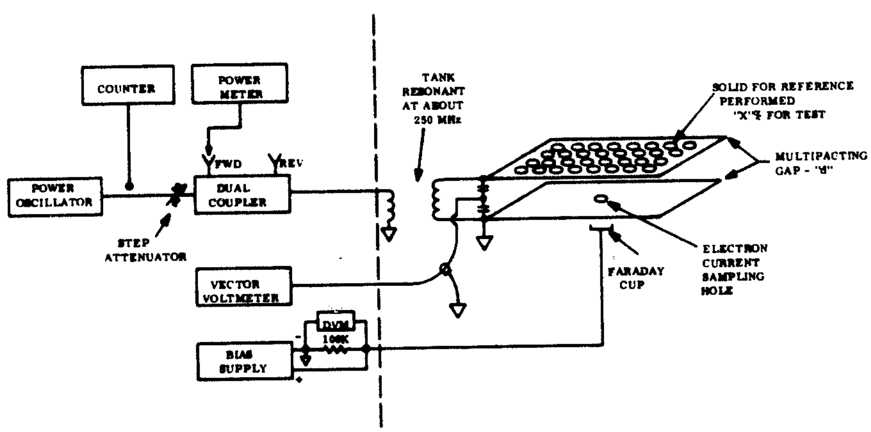


FIGURE 5-34. TEST FOR MULTIPACTOR SUPPRESSION EFFECTIVENESS OF PERFORATED SURFACES

A similar test was run with copper plates. In this case, the starting power for solid plates was 5.7 watts, increasing to 10.8 watts with one plate perforated.

These tests showed that perforation of surfaces subject to multipaction is an effective method of increasing resistance of a component to multipactor breakdown at the expense of a modest increase in conductor loss (about 20% greater attenuation than with solid conductors in this test.) Future programs should be directed toward a quantitative evaluation of this approach in relation to hole patterns, percentage of surface removed, ratio of hole diameter to electrode spacing, electrode materials, and possible augmentation effects of biased surfaces behind the perforated plate.

5.8 TRANSMITTER SYSTEM TESTS - TASK 8

5.8.1 Test Requirements

The purpose of the transmitter system tests was to provide measured operating parameters to demonstrate its operational capabilities. Included are electrical and thermal data as well as data required for a general-purpose high-power space transmitter. The initial operational tests were followed by AM-TV tests to indicate the performance capability as a TV transmitter. The test plan also included Controlled Carrier tests to demonstrate capability for achieving a high level of AM-TV performance with a restricted capacity power supply.

The tests listed in the test plan were:

1. Power, efficiency, gain, and thermal (performance tests)
2. TV Picture Quality
3. Harmonic Output
4. Controlled Carrier Operation
5. Power Supply and Transient Effects
6. Upper and Lower Sideband Attenuation (Visual channel)
7. Aural Channel Tests

A list of the test equipment required for these tests is in Appendix C. A test exciter was assembled, interconnecting items from the list of test equipment as was shown in Figure 3-34. This will provide test signals to both the visual and aural channels at 5 watt levels for each. The visual signal generator will provide the test signal formats required to ascertain the performance in accordance with the standards of EIA-RS 240.

5.8.2 Test Results

The systems tests as described in this Section were considered to be an accumulative process of the subsystem tests. With the Y2042 tube limitation as mentioned earlier, certain tests were not possible or would not contribute useful information.

However, the major operational tests are included in the text at the appropriate places, and test data will not be repeated here. Data is included in both Sections 3 and 5 in the subsections devoted to the individual components. The test procedures described below were used with the appropriate subsystems. Substantial emphasis was placed on the operational tests and the TV picture quality tests which perhaps are the most critical for a TV broadcast satellite application. The additional tests were performed as appropriate to the subsystem developments.

The following descriptions indicate the test procedures used as well as others that would be used when the 5 kW version of the transmitter is checked out.

5.8.3 Performance Tests

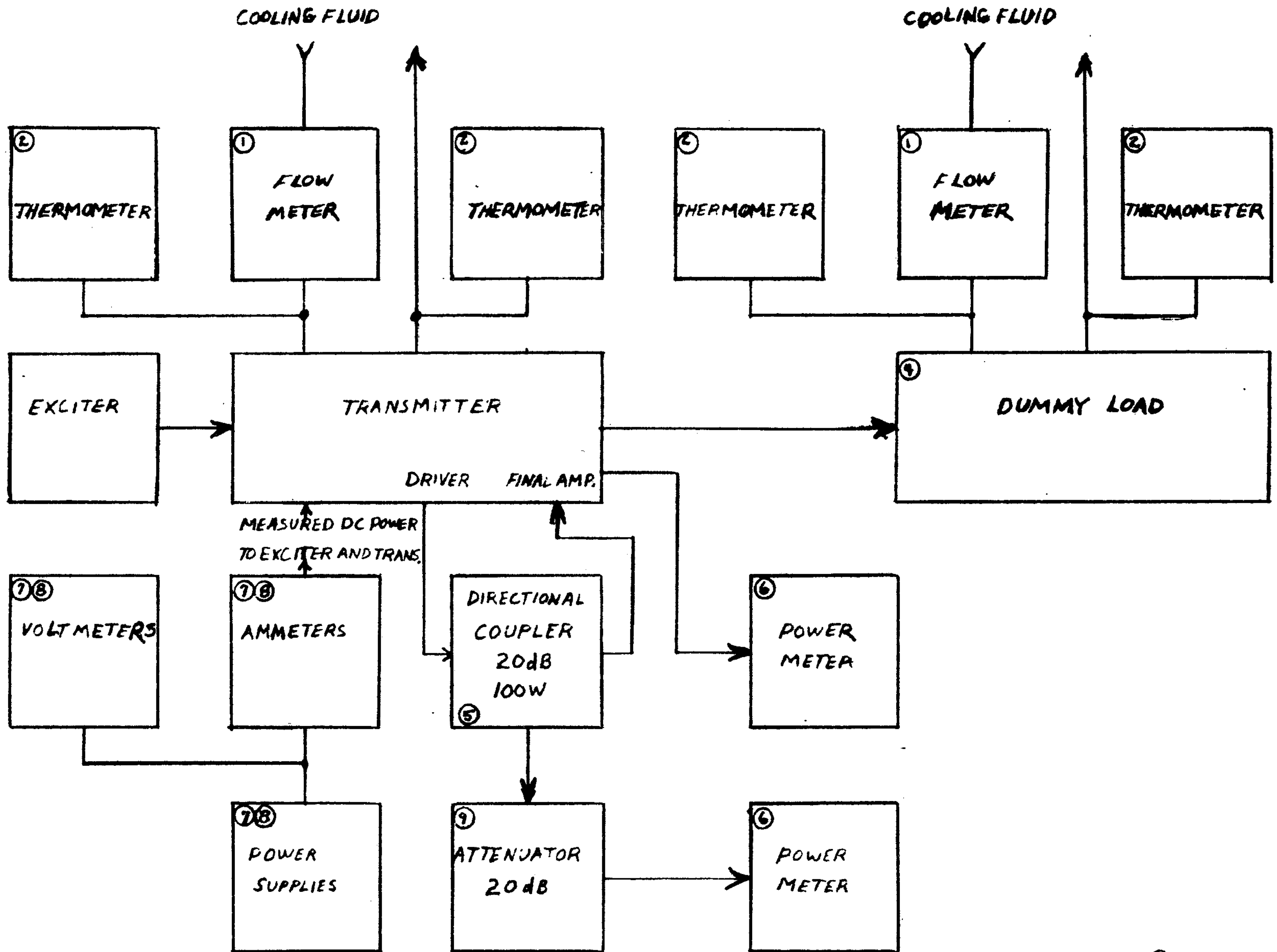
5.8.3-1 Purpose of Tests

This test series was to determine the power capabilities of the transmitter, amplifier efficiencies and gains, and heat dissipation for each portion of the equipment under various operating conditions. Tests will be made at the carrier frequency, and with necessary modulations to demonstrate operation. Unmodulated operation will be used for lower signal levels, and pulsed operation where high signal levels are involved, such as making sync peak measurements where continuous unmodulated signals would exceed tube and power supply ratings.

5.8.3-2 Description of Test

The equipment setup for the test is shown in Figure 5-35. The transmitter is temporarily modified by inserting a directional coupler between the driver and the final amplifier for the purpose of measuring transmitter stage gains. During the test all input voltages, currents, power levels, coolant temperatures and coolant flow rates are recorded.

The parameters are measured as the picture content to the transmitter is varied. Thus, the input power requirements, output power, stage gains, efficiency, and means of heat dissipation are established for each portion of the transmitter under various operating conditions.



(N) See Appendix C

FIGURE 5-35. TEST EQUIPMENT FOR OPERATING PARAMETERS MEASUREMENTS

5.8.3-3 Test Procedures

The tests will be performed starting with a white picture and increasing the gray level until the CW rating of the equipment is reached. At this point, the teti ing will proceed in a pulsed mode until tne peak sync level is reached. This is required because the transmitter is designed to handle an average picture content of about 1.6 kW, and it cannot accept the sync pulse (5 kW) or an all black picture (2.76 kW) on a continuous basis. Therefore, the duty cycle is reduced under pulse test conditions. In addition, when operating in a pulsed mode, the final amplifier will have a small idling current during the "off" portion of the cycle. This portion of power must be deducted from total amount to compute the true plate efficiency.

5.8.4 TV Picture Quality Tests

TV picture quality tests are based on the EIA Standard[3] and will cover six performance areas. These are described individually below.

5.8.4-1 Frequency Response Tests

This test is to establish the overall video amplitude versus frequency response to verify that it is in accordance with EIA standard RS-240. Figure 5-36 shows the major response requirements.

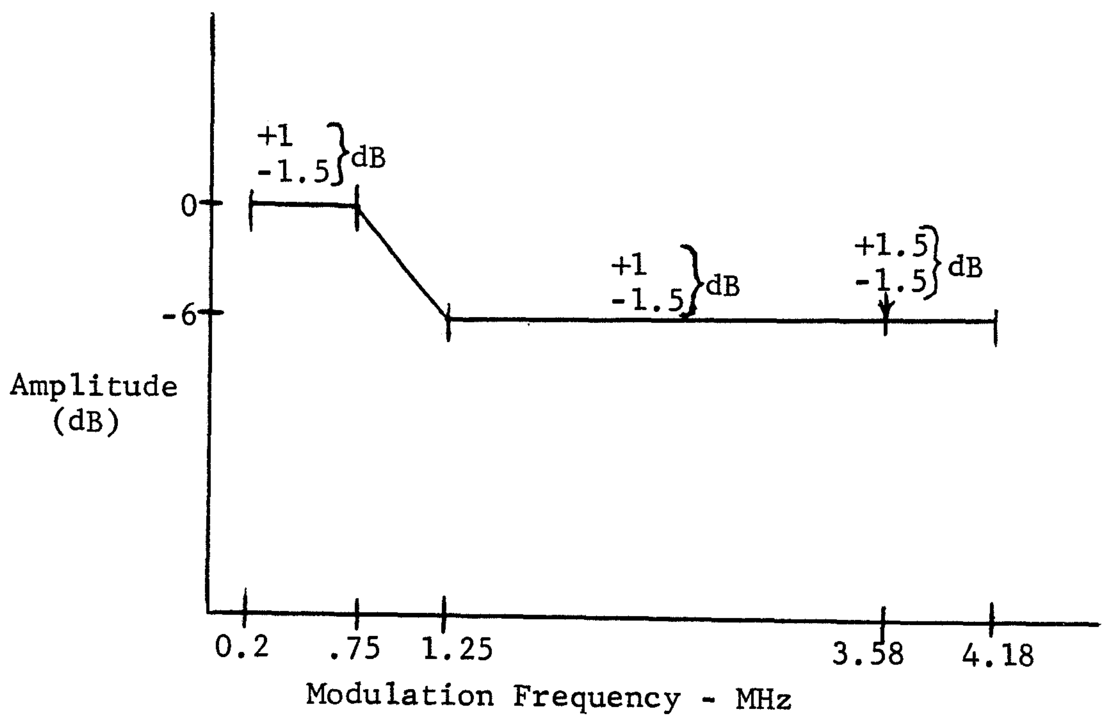


FIGURE 5-36. EIA FREQUENCY RESPONSE AND LIMITS

The equipment setup for the test is shown in Figure 5-37. An exciter tracking-receiver generates a swept video response signal. The built-in receiver portion of the test equipment, a narrow band tracking receiver synchronixed with the exciter displays the amplitude versus frequency characteristic of the transmitter under test. The diode detector is attached to the directional coupler forward power monitor terminal to measure the transmitter output response. The amplitude versus frequency response is photographically recorded from the oscilloscope.

5.8.4-2 Linearity Test

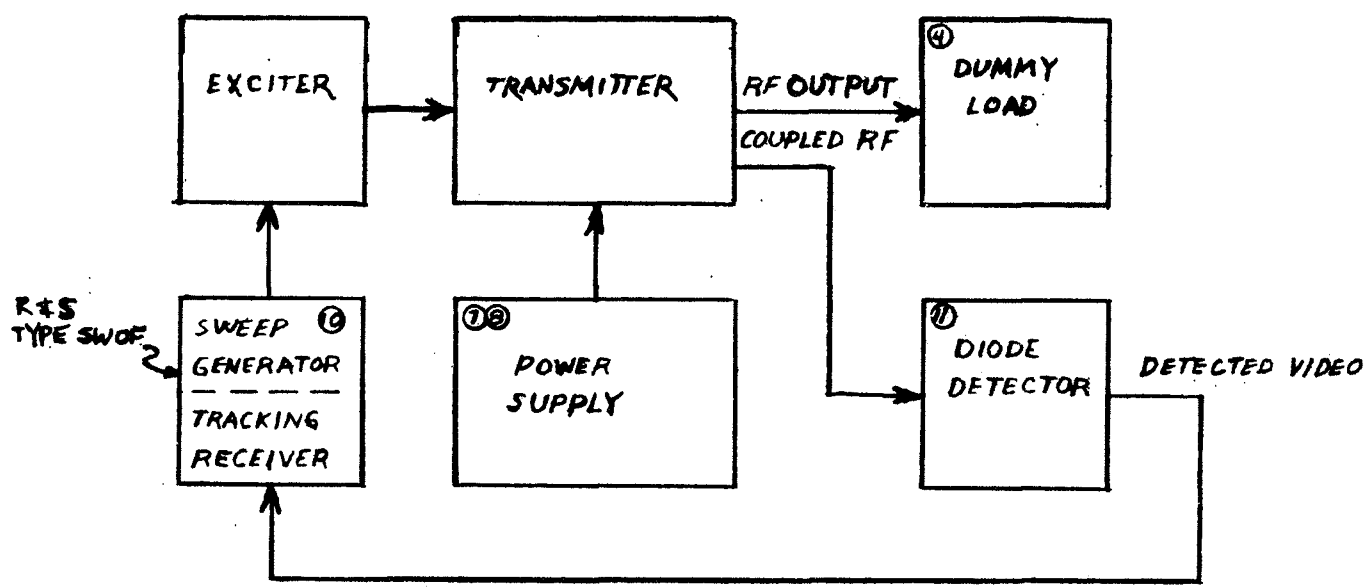
This test is to determine the low frequency (200 kHz) output amplitude versus input amplitude for the transmitter. The non-linearity shall not exceed 1.5 dB (referenced to the amplitude of the greatest step) at the 10, 50, and 90% APL (average picture level) when using a stairstep signal having ten steps of equal amplitude from the reference white to pedestal level region.

The equipment setup for the linearity test is shown in Figure 5-38. The exciter provides an AM signal to the transmitter. The signal is modulated by synchronizing pulses and a standard five or ten step video signal. The transmitter RF output is detected and fed to an oscilloscope with a type "W" preamplifier and response is recorded photographically.

5.8.4-3 Differential Gain Test

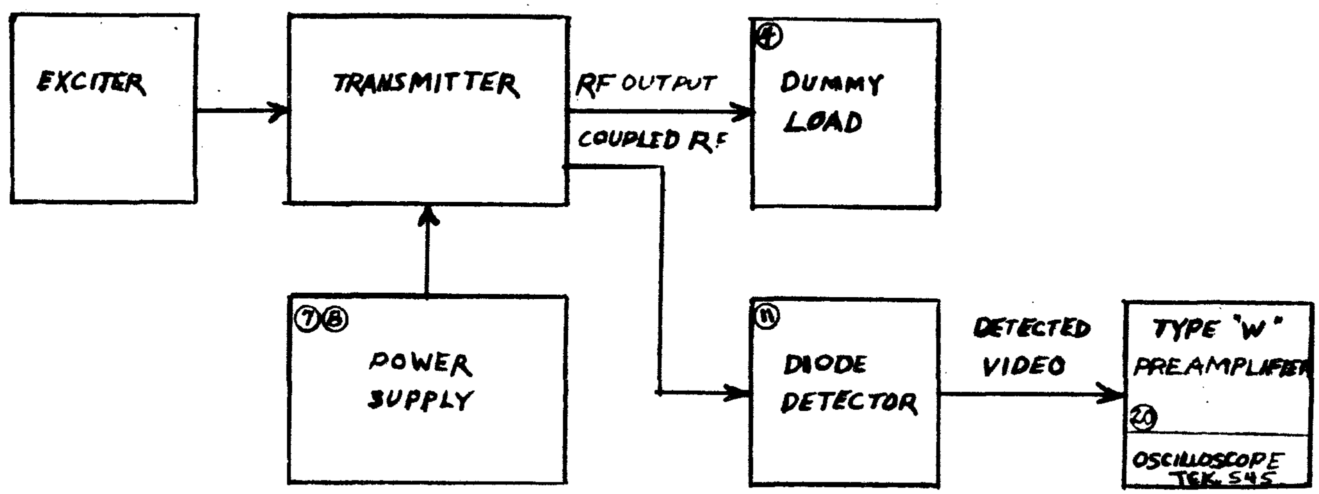
This test is to measure the differential gain of a 3.58 MHz signal as the average picture level (APL) is varied from 10 to 50 to 90%. This test requires that a 3.58 MHz sine wave super-imposed on a low frequency composite signal shall not vary more than 1.5 db for 10, 50, and 90% APL (average picture levels) signals, with the maximum gain region used as the reference.

A composite modulated signal consisting of synchronizing pulses, ten step picture level, and 3.58 MHz subcarrier is fed to the transmitter as shown in Figure 5-39. The output is video detected and the 3.58 MHz signal selected by means of a filter in the test equipment. This signal is displayed on a vectorscope (or alternately, a



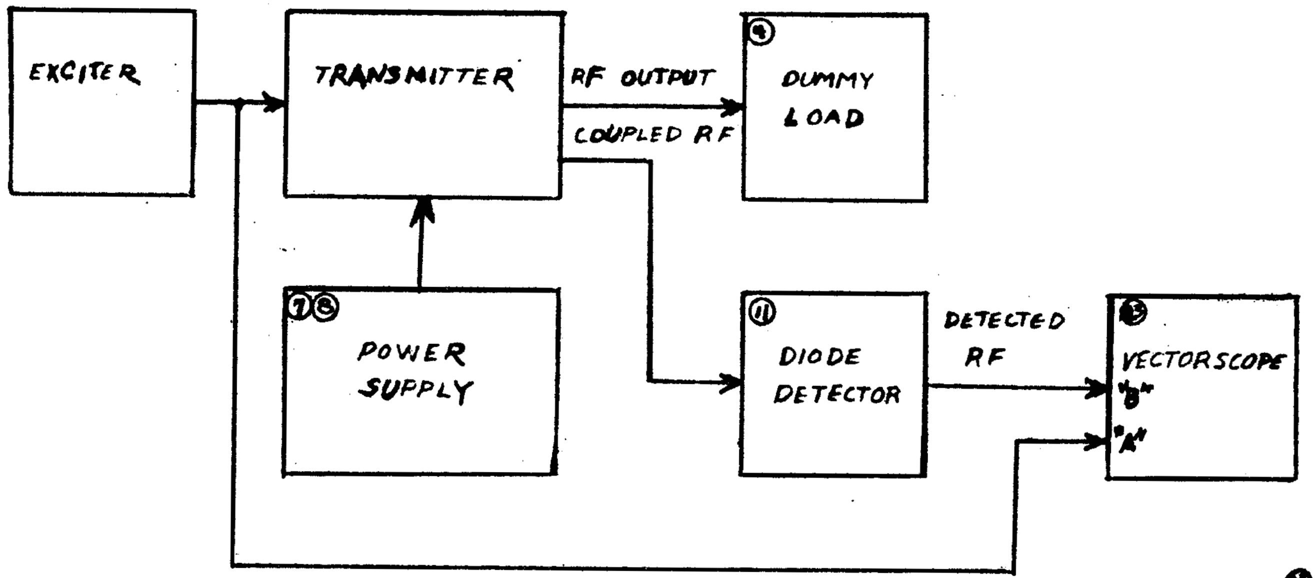
Ⓝ See Appendix C

FIGURE 5-37. TEST ARRANGEMENT FOR FREQUENCY RESPONSE TESTS



See Appendix C

FIGURE 5-38. TEST ARRANGEMENT FOR LINEARITY TESTS



See Appendix C

FIGURE 5-39. TEST ARRANGEMENT FOR DIFFERENTIAL GAIN AND PHASE MEASUREMENTS

waveform monitor) and examined for gain variations.

5.8.4-4 Differential Phase Test

This test is to measure the differential phase of a 3.58 MHz signal as the average picture level (APL) is varied from 10 to 50% to 90%. The test will verify that the differential phase is less than $\pm$ 7 degrees at 3.58 MHz when using the burst region as reference. In addition, the total differential phase between any two brightness levels will not exceed 10 degrees.

In the test arrangement of Figure 5-39, the visual transmitter input terminal will be fed a composite signal consisting of synchronizing pulses and a low frequency signal with a superimposed 3.58 MHz sine wave signal having a peak to peak amplitude of 20% of the low frequency signal amplitude between blanking and reference white. This composite test signal will be sufficient to modulate the visual transmitter to reference white while maintaining rated blanking level and rated visual transmitter output power. The coupled RF output is video detected and the 3.58 MHz signal extracted by means of a filter in the vectorscope. The phase of each step is compared in phase to that in the burst region.

5.8.4-5 Envelope Delay

This test is to measure the envelope (group) delay versus frequency characteristic of the visual transmitter. The envelope delay is defined to be the first derivative of phase with respect to angular velocity. The standard is portrayed graphically in Figure 5-40.

The equipment and setup for the test is shown in Figure 5-41. A swept frequency generator is modulated by the envelope delay measuring unit (EDMU); this signal in turn modulates the visual exciter. The transmitter output is detected and then demodulated by the EDMU. The demodulated signal is then compared with the original signal for the actual envelope delay.

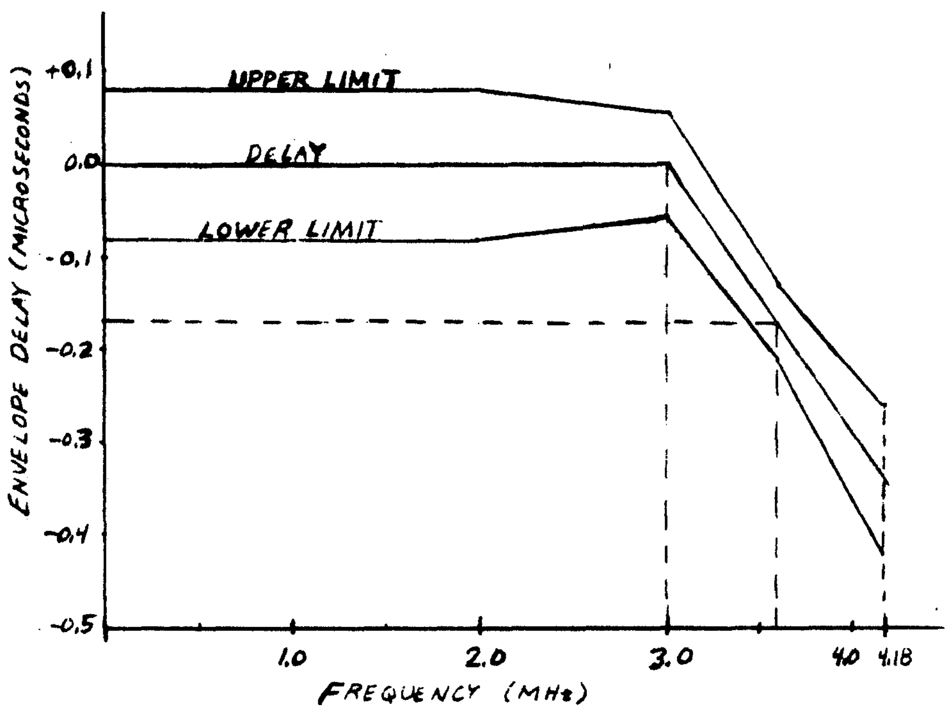


FIGURE 5-40. ENVELOPE DELAY FOR COLOR TV

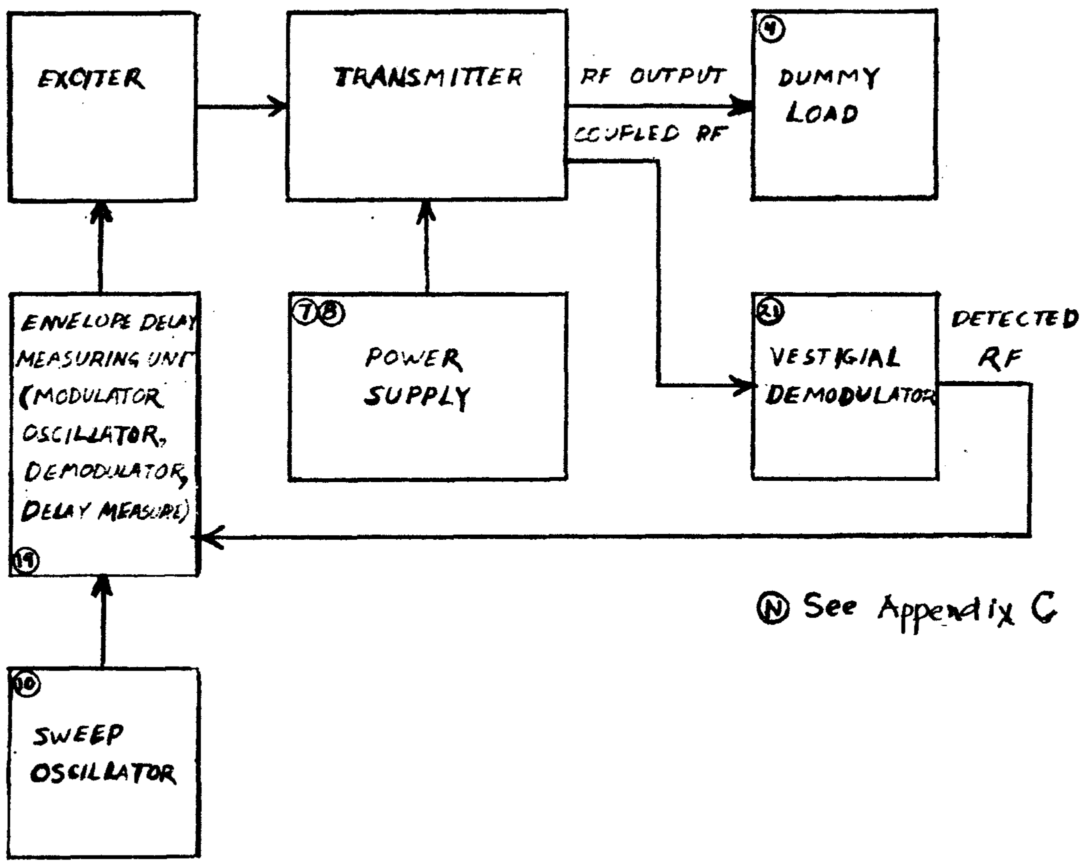


FIGURE 5-41. TEST ARRANGEMENT FOR ENVELOPE DELAY MEASUREMENT

5.8.4-6 Hum and Noise

This test is to measure the hum and noise modulation in the amplitude of the RF output that is not produced by the video modulation signal. The hum and noise level within a band of 30-15,000 Hertz should be at least 40 db below the level which would be produced by 100% modulation of the transmitter with a single frequency sine wave where 100% modulation is defined as the synchronizing pulse peak level.

The test equipment is set up as shown in Figure 5-42. No modulation is applied to the video input so that the only modulation in the output is that induced by hum and noise generated in the equipment. The low pass filter removes noise and hum above 15 KHz. The hum and noise in the output is detected on an oscilloscope.

5.8.5 Harmonic and Spurious Outputs

The purpose of this test is to measure all harmonic, subharmonic and spurious radiation from the transmitter as an indication of filtering required in future transmitters of similar design to insure that these radiations are at least 60 dB down from the peak visual carrier. A probe is inserted into the output waveguide to sample the RF output as is shown in the equipment block diagram of Figure 5-43. The sampled RF signal then enters the band reject filter prior to the spectrum analyzer. The band reject filter suppresses the carrier and permits the measurement of the low level harmonics and spurious signals.

Testing will be performed by driving the exciter at normal black level, with color subcarrier sync signals. The band reject filter is detuned and the variable attenuator adjusted so as to achieve a suitable reference level for the carrier peak visual power. The band reject filter is then adjusted to suppress the carrier and prevent overdriving the spectrum analyzer. The attenuation between the probe and the analyzer is reduced in calibrated amounts and the relative levels of subharmonics, harmonics, and spurious signals measured. Levels from 600 MHz to 10 GHz are recorded except the visual passband of the transmitter is omitted.

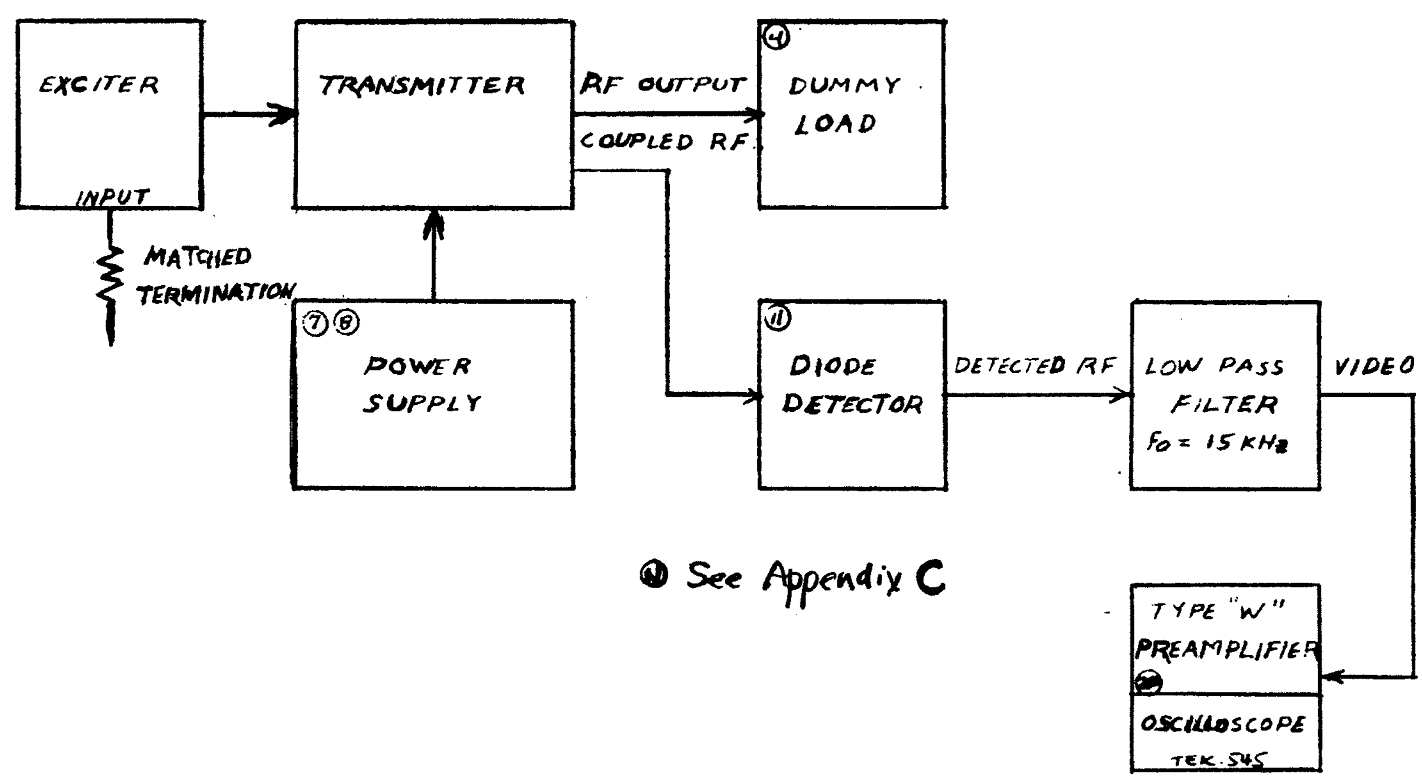


FIGURE 5-42. TEST ARRANGEMENT FOR HUM AND NOISE MEASUREMENT

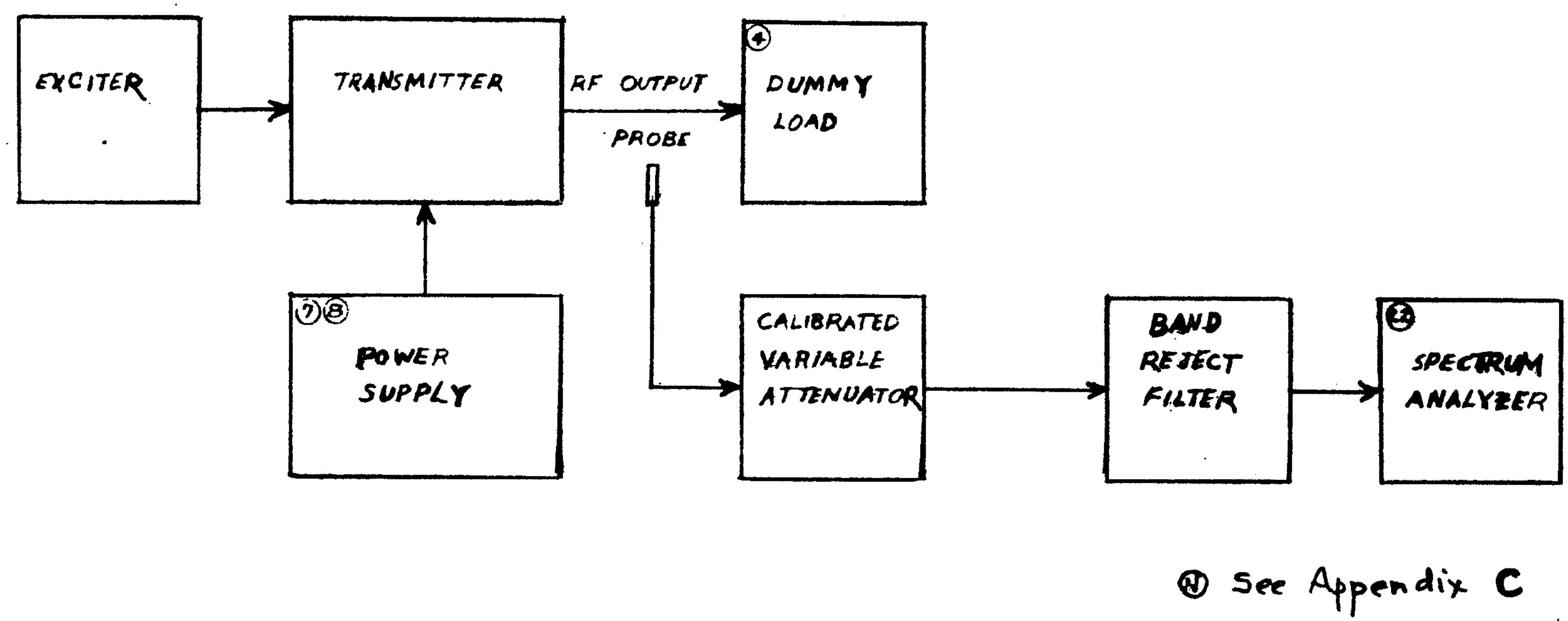


FIGURE 5-43. TEST ARRANGEMENT FOR HARMONIC AND SPURIOUS OUTPUTS MEASUREMENTS

Since the measurements are made in half height waveguide (WR 975), precautions must be observed. This waveguide will not support frequencies below 605 MHz so no measurements will be required below this frequency. This waveguide will support spurious modes above 1210 MHz; therefore, measurements with the probe must be considered as indicative rather as absolute measurements of harmonic levels. By judiciously probing the waveguide in different physical positions, the likelihood of "missing" a harmonic signal by probing into a null region will be greatly reduced. Elaborate schemes for measuring harmonic levels in waveguide use this basic multiple sample approach plus an analysis routine but this is considered outside the scope of the present study. Instead, a simple level estimate based on harmonic voltage spatial pattern, will suffice for this order of magnitude measurement.

Care must also be observed in using the probe. Spurious results can be obtained with certain probe configurations. Therefore, probes should be calibrated in a section of TEM line over the frequency range of interest to verify that valid data is obtained.

5.8.6 Controlled Carrier Operation

These tests are to evaluate the controlled carrier mode of the AM-TV transmitter operation. The measurements in these tests will be compared with those for the transmitter in the normal mode of operation, and judgements made as to the effectiveness of the controlled carrier mode.

The tests described in Section 5.8.3 are repeated for transmitter operation in the controlled carrier mode. Specifications for each test remain as for the previous tests. Specific items of interest are possible improvements in transmitter system efficiency and effects on TV picture quality when carrier control is used.

5.8.7 Power Supply Regulation and Effects of Transient Loading

This test is to evaluate the performance of the power conditioner unit and LC energy storage filter when subjected to normal transient conditions. These transient conditions would be caused by radical changes in the picture content of the visual carrier. This test will be performed with the transmitter operating in the controlled carrier mode and the non-controlled carrier mode to enable the evaluation of additional possible improvements.

The transmitter and test equipment will be set up as shown in Figure 5-44. A pulse generator drives the exciter and provides a transition in the APL during a frame interval. Thus, the picture level changes abruptly from one level to another. This provides the transient load on the power supply, power conditioner, and LC filter. The oscilloscope is synchronized to the pulse generator and displays the voltage waveforms at the required points in the system. Thus, any effects of the transient will be observed in the oscilloscope waveforms.

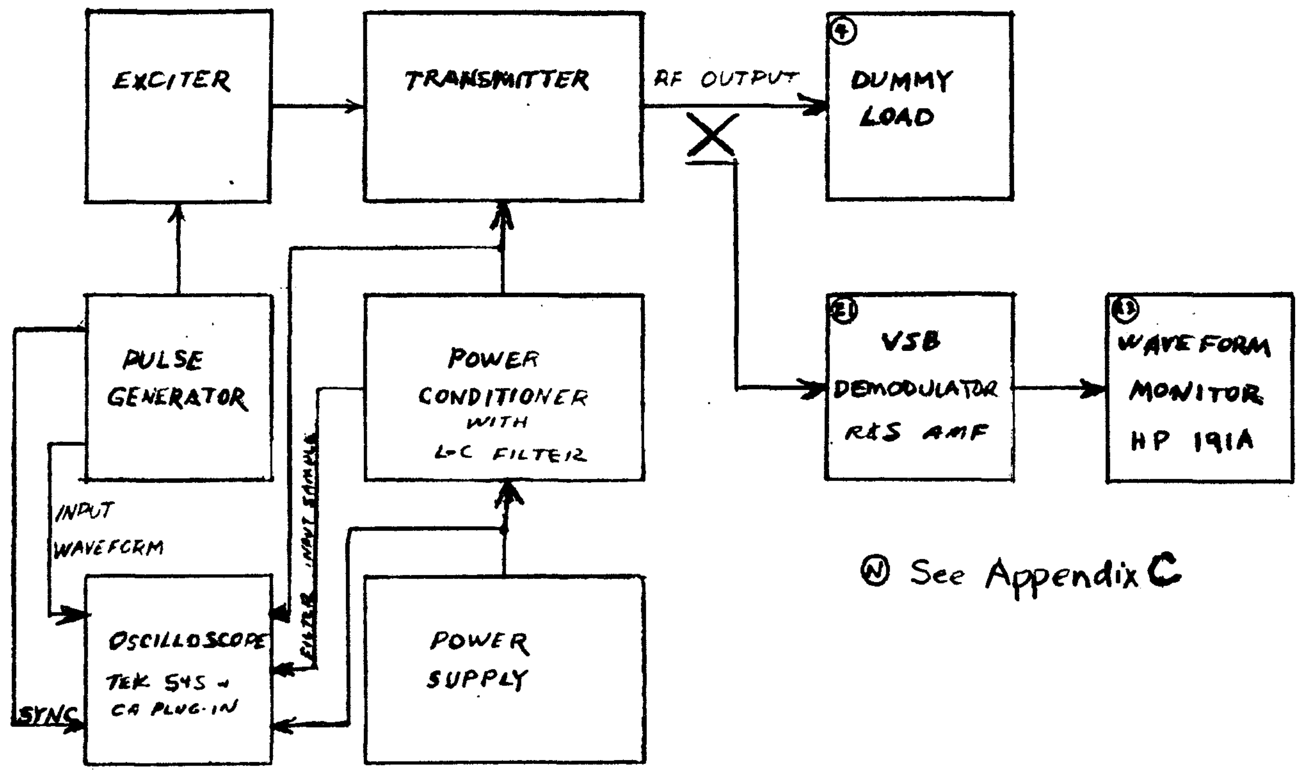


FIGURE 5-44. TEST ARRANGEMENT FOR EFFECTS OF TRANSIENT LOADING ON THE POWER SUPPLY

5.8.8 Upper and Lower Sideband Attenuation in the Visual RF Channel with VSB and Color Image Filters

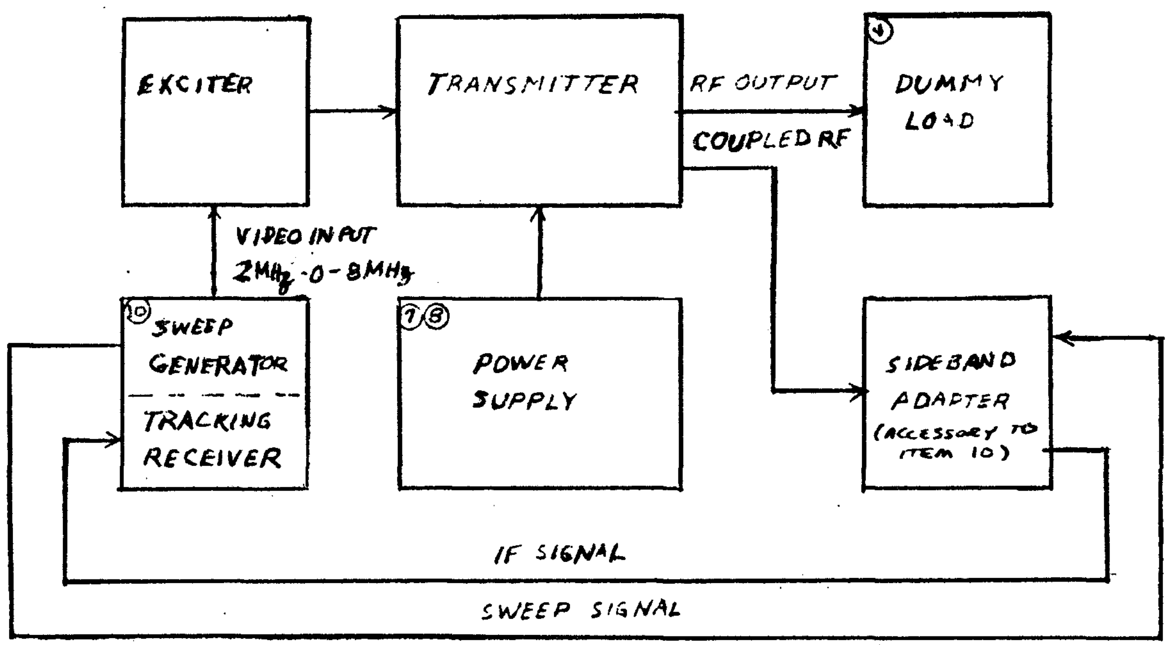
This test is to measure the upper and lower sideband attenuation and verify that the transmitter meets bandpass specifications. The lower sidebands are required to be not greater than -20 dB from the reference at 200 KHz. The color subcarrier image is required to be not greater than -42 dB from the reference. The lower sideband is defined as -4.25 to -1.25 MHz from video carrier and the upper sideband is defined to be from +4.75 MHz to +7.75 MHz from the video carrier.

The test to be performed is similar to that described in Section 5.8.4. That test measured the video passband response whereas this test measures the RF response in both the passband and the sidebands. The equipment set up is shown in Figure 5-45.

The sweep generator drives the exciter with a sawtooth from 0 to 8 MHz, with a test amplitude of 0.45 of the voltage of the synchronizing pulse peaks. The output of the transmitter is doubly converted down in the sideband adapter and fed into the tracking receiver. Although the transmitter carrier has asymmetrical sidebands, the tracking receiver measures only a 1 KHz slot at an instant of time and thus gives true sideband response of the equipment.

5.8.9 Aural Channel Amplifier

The purpose of this test is to measure the input power requirements, output power, efficiency, heat dissipation, and mode of heat dissipation for the transmitter. The equipment set-up for the test is included in Figure 5-35. The transmitter is temporarily modified by inserting a directional coupler between the driver and the final amplifier for the purpose of measuring transmitter stage gains. During the test all input voltages, currents, power levels, coolant temperatures, and coolant flow rates are recorded. The input power requirements, output power, stage gains, efficiency, and means of heat dissipation are established for each portion of the transmitter under operating conditions.



See Appendix C

FIGURE 5-45. TEST ARRANGEMENT FOR SIDEBAND ATTENUATION MEASUREMENTS

The equipment is set up as shown in Figure 5-35 and the transmitter tuned for full output. The currents, voltages, powers, coolant temperatures and coolant flow rates are recorded. From these data, the transmitter efficiencies and heat dissipation factors are computed.

In addition, a test will be performed to measure the bandwidth characteristics of the aural transmitter. This test will measure the passband flatness and indicate possible degradation of the audio signal amplified in the transmitter.

The equipment for the test is set up as shown in Figure 5-46. A sweep oscillator scans the passband of the transmitter, and a portion of the transmitter rf is coupled off and the transmitter passband is measured on the network analyzer. The passband is recorded photographically.

A third aural channel test will measure the hum and noise modulation in the amplitude of the output of the transmitter that is contributed by the transmitter. The transmitter noise should be at least 50 dB below 100% amplitude modulation within the band of 50 to 15,000 Hz.

For these hum and noise tests, the AM noise and hum of the exciter will be measured first, and then the AM noise and hum of the entire transmitter measured. A comparison of the results will indicate the contribution of the transmitter alone.

The equipment is connected first as shown in Figure 5-47 to determine oscillator noise. No modulation is used so that the measured AM noise and hum are those of the exciter. Now, connecting the equipment as shown in Figure 5-42, the noise and hum of the entire transmitter is measured. The data of the two tests will permit a determination of the hum and noise induced by the transmitter.

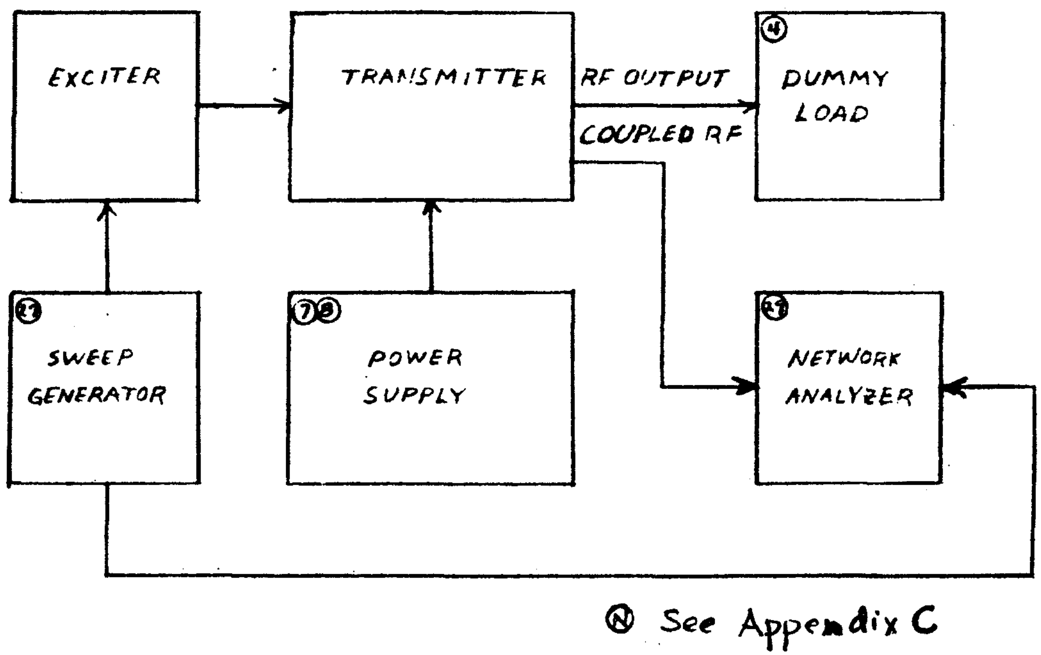


FIGURE 5-46. TEST EQUIPMENT FOR AURAL CHANNEL BANDWIDTH MEASUREMENTS

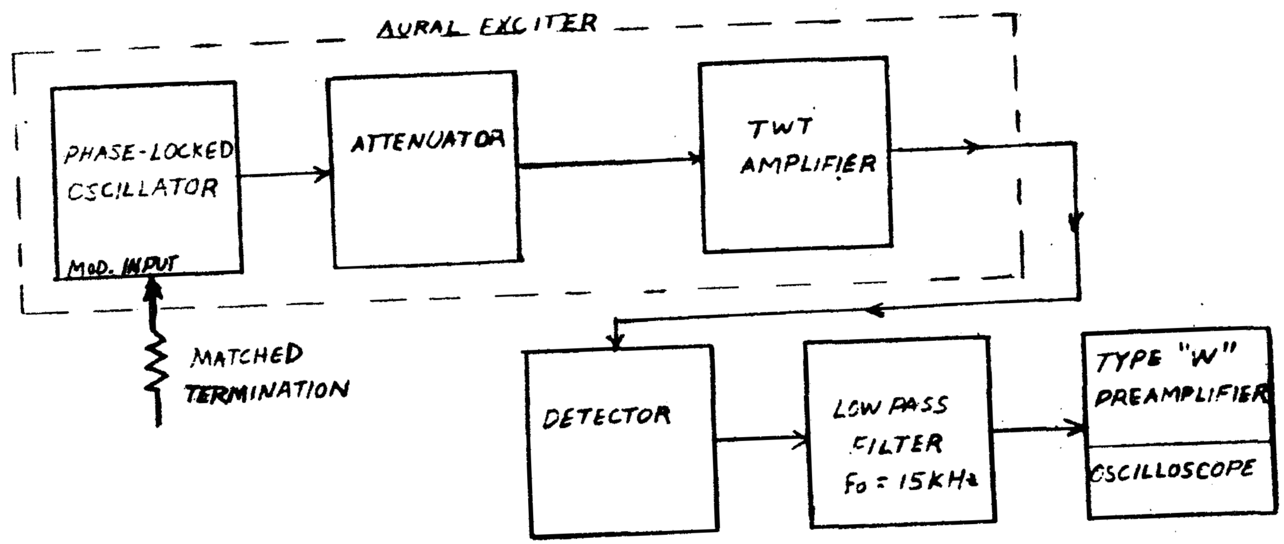


FIGURE 5-47. TEST EQUIPMENT FOR THE AURAL CHANNEL HUM AND NOISE MEASUREMENT

REFERENCES

1 - Multikilowatt Transmitter Study for Space Communications Satellites, Volume II, Final Technical Report, Phase I, General Electric Company, Philadelphia, report No. 68SD4268, Contract NAS8-21886, for NASA/MSFC, 10 June 1968.

2 - Multikilowatt Transmitter Study for Space Communications Satellites, Volume II, Final Technical Report, Phase II, General Electric Company, Philadelphia, Contract NAS8-21886, for NASA/MSFC, 30 July 1969.

3 - EIA Standard RS-240, Electrical Performance Standards for Television Broadcast Transmitters, April 1961.

4 - Tube Performance Computer, EIMAC Application Note Nr. 5, Copyright 1952 and 1967 with Addendum Copyright 1964.

5 - D. L. Smith, Mathematical Analysis of Envelope Delay in the Visual Pass Band of a 50kW Vestigial Sideband Filter, Masters Thesis, Syracuse University, 1957.

6 - Dr. K. Tomiyasu, On High Power Testing - II, General Electric Internal Memorandum, PIR #U-1J40-MKTS-011, 24 November, 1969.

7 - A. J. Geia and Dr. M. C. Jeruchim, An Experimental Investigation of the Effects of Controlled - Carrier on Television Picture Quality, General Electric Company, Report No. 69SD224, 14 March 1969.

8 - S. Ramo and J. R. Whinnery, Fields and Waves in Modern Radio, John Wiley and Sons, 1953

9 - G. L. Matthaei, L. Young, and E.M.T. Jones, Microwave Filters, Impedance Matching Networks, and Coupling Structures, McGraw Hill, New York, 1964.

10 - H. J. Riblet, The Short-Slot Hybrid Junction, IRE Proceedings, Vol. 40, No. 2, pp 180-184, February 1952.

11 - A. Viestra and R. C. Butman, Secondary Emission Characteristics of High Power Microwave Tube Surfacer, Lincoln Laboratory, MIT Technical Report No. 257, February 1962.

12 - Hesselbacher, R.W., et al, Television Broadcast Satellite (TVBS) Study-TVBS Technical Report, Vol. III, Contract NAS 3-9708, for NASA Lewis Research Center, Cleveland, Ohio, by General Electric Company, Space Division, Philadelphia, Pa. 15 November 1969.

APPENDIX A

SUMMARY OF CONSTRAINTS ON PROGRAM

The following constraints were included in the contract specifications, or appeared implicitly from the evolution of the study, EIA standards such as RS-240, and available state-of-the-art documents.

A-1 <u>System Design Study</u> - Task 1

The contract specifies the development and initial breadboarding of the stages and subsystems required for a high-power, high efficiency UHF television transmitter breadboard which can lead to the preliminary design of a multikilowatt transmitter for space applications. Program task scheduling was arranged so that all hardware design and development task requirements were defined in this task, including:

- Visual Chain Amplifiers - Task 2a and 2b
  - 50 kW Doherty Output Stage
  - 50 W Class B Linear Driver
- Aural Stage Amplifier - Task 3
  - 500 W Class C FM Amplifier
- RF Components - Task 4
  - Color Image Filter
  - RF Transmission Elements
  - Vestigial Sideband (VSB) Filter
- Monitoring and Protective Circuit - Task 5
- Controlled Carrier Design - Task 6

In addition, certain assumptions and definitions were made to allow design of the subsystems to be firmly defined:

1. EIA Standard RS-240 was used as a guide to transmitter performance requirements and test methods.
2. UHF Television channel 73 (825.25 MHz) was selected.
3. The design of the transmitter should not preclude later integration with heat pipe and power conditioner systems which may be developed elsewhere.

4. Techniques applicable to, and required for, space operation were used wherever feasible within the limits of present technology and of contract funding and schedule. Otherwise, space design requirements and approaches which might be used in future designs are indicated and discussed where applicable.

The basic approaches and specifications for the transmitter subsystems were determined in this task. The approach used was:

1. Employ previous MKTS and other related study results as a guide in establishing transmitter design approaches and as sources for design data. Determine preferred tubes, transmission line types, RF amplifier specifications, other circuit requirements, and supporting equipment requirements.

2. Determine the availability of major components required for use in the transmitter breadboard.

3. Perform design and tradeoff studies as required to arrive at an optimum design. Consider:
   a) Incorporation of space-required features where feasible.
   b) High-efficiency
   c) High quality signal channel performance

4. Specify or otherwise define subsystem requirements which will result in the attainment of performance objectives and a well-integrated breadboard test unit. Implement state-of-the-art advances as feasible.

A-2 Visual Channel Amplifiers - Task 2

A-2.1 Driver Amplifier

Development of a linear rf amplifier suitable for incorporation into the TV transmitter breadboard visual amplifier chain was required as the task output. Design objectives for the amplifier were:

Electrical

Operating Frequency	825.25 MHz (video carrier) (tunable cavity designed to permit tube change)
Bandwidth	824.0 to 829.5 MHz at $\pm$ 0.5 dB (min. BW)
Power Output	125 watts sync peak (capability for 200 W sync peak would be desirable)

Power Input Level	5 watts sync peak (max.)
Gain Variation with Drive Signal Level	0.5 dB (maximum); linearity should be adequate to meet EIA standards
Phase Transfer Variation with Drive Signal Level	$\pm 3^{o}$ (maximum)
Efficiency	>50% at rated output into a matched load

<u>Thermal</u>

Cooling Method	Conduction or radiation (No forced air)
Heat Sink Temperature	100°C (max.)
Maximum Tube Seal Temp.	250°C Maximum

<u>Mechanical</u>

Cavity Construction	Breadboard design should be adaptable to space-type hardware with minimal changes. Design features should include: Ruggedness Avoidance of excessive mechanical stresses on the tube and other amplifier components, including thermal expansion. Cavity should be capable of being readily dismantled for tube replacement, developmental changes, etc.
Compatibility with Breadboard Circuit	Designs should be compatible with all electrical and mechanical interfaces in the breadboard circuit.

<u>Personnel Safety</u> - Must include considerations of:

High Voltage, rf radiation and external hot spot temperature.

A-2.2 Doherty High Efficiency Amplifier Design

Design of gridded tube rf amplifier cavities and associated circuitry in the Doherty configuration, suitable for incorporation into the MKTS III TV transmitter breadboard video amplifier chain, is required. Design objectives for the Doherty amplifier were:

Electrical

Operating Frequency	825.25 MHz (video carrier)
Bandwidth	824.0 to 829.5 MHz at $\pm$ 0.5 dB (Min. BW)
Power Output	5.0 Kilowatts peak sync
Power Input Level	65 watts sync peak (nominal) at the input terminal of the rf input power divider
Gain Variation with Drive with Drive Signal Level	1.0 dB (maximum) - dynamic bias is required to achieve this
Phase Transfer Variation with Drive Signal Level	$\pm 3^{o}$ (maximum)
Efficiency Objective	$\geq$ 60% at video signal level of 25% of rated output power and $\geq$ 65% at rated peak sync power into a matched load

In addition, thermal control was required, using water cooling for the high power stages in the transmitter tests, and maintaining temperatures no greater than the following:

Heat Sink Temperature	60^{o}C (max.) for cooling water 100^{o}C (max.) for cavity elements 300^{o}C (max.) for the conduction cooled anode
ΔT Between Adjacent Tube Seals	100^{o}C (maximum)
Maximum Tube Seal Temp.	150^{o}C maximum at the grid seal. Other seals must observe the ΔT limit above.

Mechanical design and personnel safety were considered in the configuration used in the final tests, as outlined for the driver stage.

A-3 Aural Channel Amplifier - Task 3

Development of a triode rf amplifier suitable for incorporation into the transmitter breadboard was required. Design factors were:

Electrical

Operating Frequency	829.75 MHz (aural carrier)
Bandwidth	100 KHz at $\pm$ 0.5 dB (min. BW)

Power Output	500 watts CW
Input Power Drive Level	5 watts CW (max.)
Efficiency Objective	≥ 70% at rated output into a matched load (design goal)
Circuit Configuration	DC grounded anode

Thermal

Water cooled anode, others conduction or radiation cooled (no forced air), $100^{\circ}C$ heat sink, tube temperatures the same as in Section A-2.2 (Doherty amplifier).

Mechanical

Should have no features which preclude adaptation to a space qualified transmitter. Basically the same as for the Doherty amplifier, Section A-2.2

Personnel Safety

To be observed at all times with respect to high voltage, rf radiation, and external temperatures.

A-4 RF Components - Task 4

A-4.1 High Power Waveguide Components

This task involved the design, fabrication, and basic tests on the several rf components required in the transmitter test. The high power components are of half-height WR975 waveguide. Constraints were:

Color Image Filter	821.67 MHz 20 dB attenuation (goal, 15 dB acceptable) .05 dB Insertion Loss (goal) 1.10 max. VSWR (visual channel) mechanically and thermally compatible with rest of transmitter and system.
Hybrid, 3 dB	824 to 830 MHz sidewall coupling, 3 ± .15 dB 30 dB isolation 0.1 dB max. loss 1.05 VSWR (goal)

5.5 kW rating
Mechanically and thermally compatible with rest of transmitter and system.

Transition, half-height WR975 to 1-5/8" coax

1-5/8" line at 50 ohms
824 to 830 MHz
1.03 max. VSWR
0.1 dB insertion loss
Mechanically and thermally compatible with rest of transmitter and system.

Directional Couplers

-30 dB coupling for power monitoring
Require forward and reverse power measurement.

A-4.2 Vestigial Sideband Filter

This filter shapes the TV transmitted spectrum to conform to that of EIA Standard RS-240. An input type preceding the driver was selected since the linear amplifiers should not generate significant distortion to disturb the spectral response at the output. Thus, the filter will operate at inputs up to 10 Watts from the external exciter. Other design factors were:

824 to 830 MHz
20 dB minimum attenuation of unwanted lower sideband components
3 dB loss (max.) over pass-band
1.5 VSWR over pass-band
Response per EIA Standard RS-240 as in Figure A-1 (lower sideband shaping)

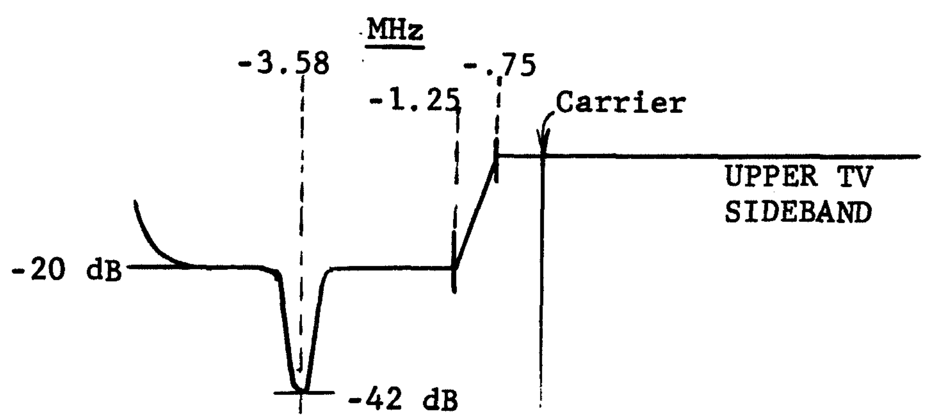


Figure A-1. Vestigial Sideband Response

A-5 Monitor and Protective Circuitry - Task 5

This task involved the design, fabrication, and testing of all monitor and protective circuits for the complete transmitter. The extent of each is as follows:

- Crowbar for Visual Channel Amplifier - the circuit reacts upon receipt of an excessive-current fault signal in the high efficiency amplifier anode circuit. A triggered spark gap (low impedance current path) is provided to divert the fault current from the amplifier such that damage to amplifier components, particularly the high power tubes, will be minimized. Typically, fault energy should be maintained below 5 joules in a tube arc situation.

- Fault Sensing and Control Logic - Provision was made for the circuit logic necessary for protection of visual and aural amplifiers and associated components, including the crowbar circuit described above. This protective and control logic can be integrated as necessary with the associated laboratory test circuit items such as power supplies, cooling system, and dummy loads.

- Monitoring - Monitoring points giving the necessary output signals required for transmitter monitoring by the associated test equipment were provided. Parameters monitored included:

  (a) Input to output signal transfer characteristics
  Gain
  Phase
  Distortion

  (b) DC to RF Power Conversion Efficiency

- Other Factors - Designs and hardware generated in this task are generally adaptable as space-type circuits and hardware. Personnel safety provisions were given due consideration in the design and testing efforts.

A-6 Controlled Carrier Circuitry - Task 6

Requirements placed on the Controlled Carrier circuitry were:

824 to 830 MHz
10 watts RF power at 60% duty factor (input to Driver)
1.0 dB or less insertion loss
Variable attenuation to 6.0 dB
Time constant = 0.2 to 20 milliseconds
1.2 VSWR goal

The control is located between the external exciter and the input terminal of the driver stage; the controlling signal is derived from the average plate current requirement. A selectable gray-level clamp is included to permit level adjustment for determining the amount of carrier control to be employed.

A-7 High Power RF Component Environmental Tests - Task 7

The contract requires the environmental testing of several RF components which include in descending order of significance, the following items.

	Component	Size
1.	3-1/8 Inch Coaxial Line Section	
2.	Half Height WR975 Waveguide Section	
3.	Coaxial Step Section, 3-1/8" Dia.	Wavelength plus 14"
4.	Waveguide Step Section, WR975	10" x 20"
5.	3 dB Sidewall Coupler, Dual WR975 at 1/2 height	20" x 24"
6.	Aural Notch Cavity Filter, WR975 at 1/2 height	10" x 14"

A program plan developed in this task was to be the basis for testing these items. Available facilities were to be used in order to achieve maximum results with a minimum expenditure. Tests on items 3 and 4 above would provide breakdown calibration points; tests on the other components would confirm their suitability for space.

A-8 Transmitter Tests - Task 8

The transmitter tests are for the complete system including all the elements generated in the separate tasks of the contract. The test plan was the basis for system tests performed. The tests were set up in terms of test equipment required, how the equipment is assembled, the test procedure, and data expected. The tests divided into the functional tests of power, efficiency, and gain measurements, followed by tests on TV performance in accordance with EIA Standard RS-240. The requirements of the latter were included in the specifications above for each task.

APPENDIX B

PREVENTION OF MULTIPACTOR BREAKDOWN BY USE OF LOW IMPEDANCE COAXIAL LINE[6]

In order to reduce the probability of multipactor occurrence, it appears worthwhile to consider the use of low characteristic impedance lines. The following analysis applies to coaxial line; however, a similar approach would be valid for other forms such as ridged waveguide with small ridge gap spacing. A differential radius of about one millimeter at 800 MHz is considered a limiting value for the gap. Ostensibly, coaxial lines can be used in resonant filters.

If the outer conductor diamter is 1.625" and the gap is 1 mm, then the inner diameter is 1.545". The characteristic impedance is

$$\begin{aligned} Z_c &= 60 \ln (D_o/D_i) \\ &= 60 \ln (1.052) \\ &= 3.04 \text{ ohms.} \end{aligned}$$

Breakdown in air occurs with a field strength of about 30,000 V/cm so that a 1 mm gap can handle about 1.5 kV with a safety factor of two. Thus, the power handling capability of this line is

$$P = V^2/Z_c \cong 370 \text{ kW.}$$

In high power klystrons, a maximum safe RF potential gradient in vacuum with copper electrodes is about 900 kV/inch; for conservative operation a gradient of 600 kV/inch is desired. With surfaces other than copper, such as molybdenum or tungsten, the permissible gradients are higher since the materials are less apt to produce surface cracks under operation. Assuming a peak value of 600 kV/inch, a safe vacuum breakdown average power level in a 1 mm gap, 3.04 ohm, copper coaxial line is

$$\begin{aligned} P &= \left[\frac{600 \times 10^3}{25.4 \text{ mm/in.}}\right]^2 \times 1/4 \times 1/(3.04) \\ &= 47 \text{ megawatts,} \end{aligned}$$

which includes a 2:1 safety factor. The attenuation of a matched 3.04 ohm coaxial line can be computed from the following:

$$\alpha = \frac{179.5}{Z_c} \times 10^{-9} \sqrt{f} \ (1/a + 1/b)$$

where

f = 800 MHz

a = inner conductor radius = 0.772"

b = outer conductor radius = 0.812"

Thus, the matched attenuation at 800 MHz is

$\alpha = 4.23 \times 10^{-3}$ dB/inch

= 0.0625 dB/wavelength

= 1.5% loss/wavelength

Appendix C (cont.)

Item	Quantity	Description	Mfgr.	Model #
18	1	Vectorscope	Tektronics	520
19	1	Group-Delay Measuring Equipment	Rohde & Schwarz	LFM
20	1	Oscilloscope Preamplifier	Tektronics	"W"
21	1	Vestigal Demodulator	Rohde & Schwarz	AMF
22	1	Spectrum Analyzer	Hewlett Packard	8551
23	1	TV Waveform Monitor	HP	191A
24	1	TV Sync Generator	RCA	TG-3
25	1	TV Color Bar Generator	RCA	TG-4
26	1	Color Broadcast Monitor	RCA	
27	1	Sweep Generator	HP	
28	1	Counter	HP	5246/5254
29	1	Network Analyzer	HP	8410
30	1	True RMS Voltmeter	Ballantine	323
31	1	Peak Reading, Tunable Voltmeter	Empire	NF-105
32	1	TV Monitor Receiver		
33	2	Directional Coupler, 20 dB, 500 W		
34	1	Directional Coupler 30 dB, 5 KW		
35	1	Side Band Adaptor	Rohde & Schwarz	

APPENDIX C

Required Equipment for Multikilowatt Transmitter Tests

Item	Quantity	Description	Mfgr.	Model #
1	2	Coolant Flow Meter		
2	4	Coolant Temperature Meters		
3	1	TV Signal Generator	Rohde & Schwarz	SDFA
4	2	5 KW Dummy Load, 50 OHM	Bird	872
5	1	Directional Coupler, 20 dB 100 W		
6	2	Power Meter	H.P.	431
7	1	Power Supply 1.5 KV 1.0A	Sorenson	DCR 1500-1
8	1	Power Supply 5 KV @ 3.0 A	G.E.	
9	2	Attenuator 20 dB, 1 W		
10	1	Sideband Response Analyzer	Rohde & Schwarz	SWOF
11	1	Monitoring Diode		
12	1	Processing Amplifier	Grass Valley	900T, 940
13	1	Gain and Phase Adjust	Grass Valley	941
14	1	Burst Amplifier	Grass Valley	966
15	1	TV Test Signal Generator	Tektronics	140
16	1	Aural FM Generator	Fairchild	MO(L)-100 XB
17	2	TWT Amplifiers	GE	